James Forney WB2CCM
New Canaan, Conn.

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Jim Fisk W1DTY
RFD 1, Box 138
Rindge, N.H. 03461

73 Useful Transistor Circuits

Useful transistor circuits for audio and speech equipment, receivers, transmitters and test equipment.

Circuits for Audio Equipment

Direct coupled amplifiers

The direct coupled amplifier illustrated in Fig. 1 is just about as simple as possible, but provides very usable results. The collectors of the first two transistors operate at about 0.3 volts; this type of operation yields somewhat less than normal gain, but provides considerable reduction in noise produced at the input by the transistor. The biasing of the first stage is controlled by resistor R1 and because of the direct coupling between stages, indirectly controls the bias to the other two stages. Since the gain and leakage varies widely between different transistors, this resistor must be adjusted experimentally to provide optimum bias for the last transistor (Q3). This is easily done by adjusting its value until there is 0.8 volts across the headphones (points A and B in the schematic).

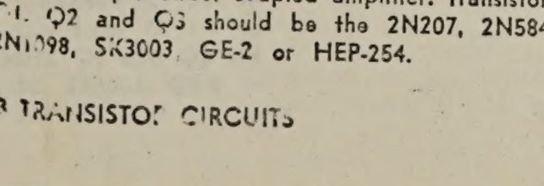
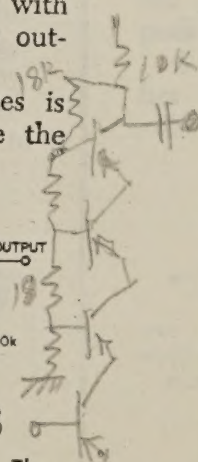


Fig. 1. Simple direct coupled amplifier. Transistors Q1, Q2 and Q3 should be the 2N207, 2N584, 2N1098, SK3003, GE-2 or HEP-254.

Since R1 is connected to the collector of Q3, bias variations caused by changes in temperature are reduced by negative dc feedback introduced by this resistor. For example, as the leakage in Q1 increases with temperature, the collector voltage on Q3 decreases. The increased leakage is partially compensated for because the lower voltage on Q3 causes less current to flow through R1. Generally speaking, this circuit will compensate quite nicely against temperature changes up to about 100°F. Above 100° it is possible that the transistor will be driven into nonlinear operation with resultant distortion and reduced power output.

The dc resistance of the headphones is very important in this circuit because the



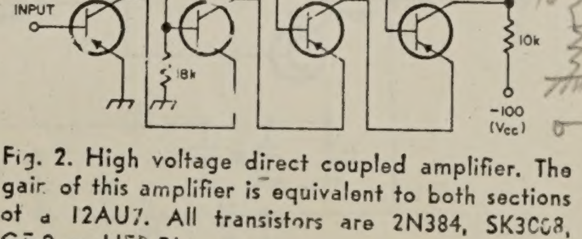


Fig. 2. High voltage direct coupled amplifier. The gain of this amplifier is equivalent to both sections of a 12AU7. All transistors are 2N384, SK3008, GE-9 or HEP-51.

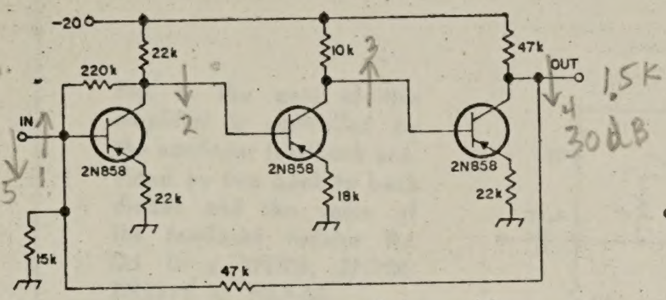


Fig. 3. This simple direct coupled amplifier provides 30 dB gain and identical 1500 ohm input and output impedances. For higher gain, similar units may be cascaded up until 10 volts peak to peak is obtained at the output.

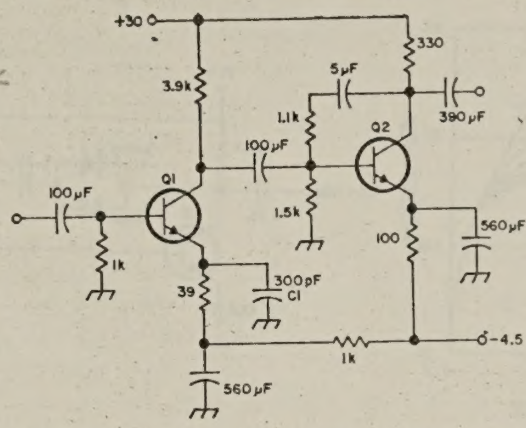


Fig. 5. This wideband amplifier exhibits 26 dB gain from 5 Hz to over 30 MHz and will deliver a 7 volt signal into a 100 ohm load. Transistors Q1 and Q2 are 2N2218.

voltage drop across them determines the operating conditions of all three stages. For optimum operation, the dc resistance of the earphones should be in the neighborhood of 1000 ohms. Most phones with an impedance of 2000 ohms have a dc resistance of 1000 ohms, but if you're in doubt, the resistance may be easily measured with an ohmmeter.

The main advantage of the high voltage direct coupled amplifier in Fig. 2 is that it may be connected directly to a rather high value of B+. Its gain is equivalent to a single 12AU7 (both sections) and because of the direct coupling, provides extremely wide bandwidth. Although the input impedance of this circuit is only 2000 ohms, it is still very useful for many applications where a simple amplifier is required.

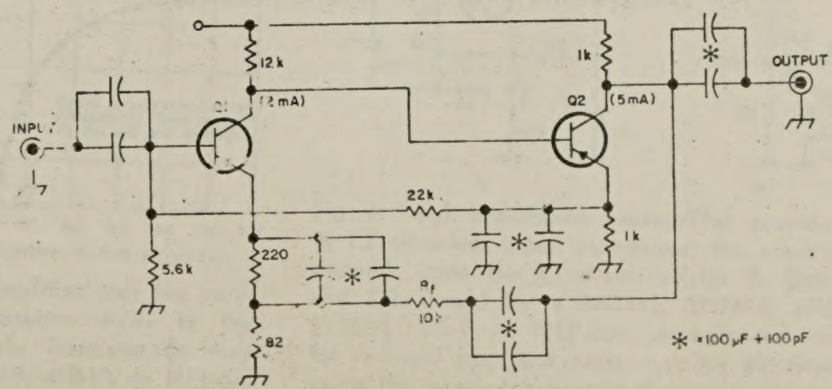
Another very simple direct coupled amplifier is illustrated in **Fig. 3.** This amplifier provides almost exactly 30 dB gain and has identical 1500 ohm input and output impedances. For extremely high gain then, similar units can be cascaded up until an output voltage of 10 volts is obtained. This amplifier is also quite wideband, and with the transistors specified, the gain is essentially flat up to about 1 MHz.

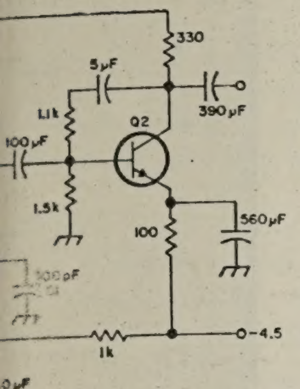
Wideband amplifiers

In the wideband amplifier shown in Fig. 4, the gain is controlled by the feedback resistor R_f. With a 10 kilohm feedback resistor, the gain is greater than 30 dB from 10 Hz to 17 MHz. When the resistor is completely removed from the circuit, the gain is greater than 50 dB up to about MHz, but the biasing of the input transistor becomes very critical to prevent signal distortion. Note that the large electrolytic coupling capacitors should be paralleled with smaller capacitors that have good high frequency characteristics.

Another wideband amplifier is illustrated in **Fig. 5**; this amplifier has a frequency response from 5 Hz to over 30 MHz. The voltage gain over this range is 26 dB and the amplifier will deliver an undistorted 7 volt sine wave into a 100 ohm load. This circuit has excellent stability and linearity, and by adjusting the bias and emitter bypass capacitor C1 experimentally, the frequency response may be increased up to 50 MHz.

Fig. 4. The gain of this wideband amplifier may be controlled by the value of the feedback resistor R_f. The 10K resistor shown here provides more than 30 dB gain from 10 Hz to 17 MHz. Q1 and Q2 are 2N2188, SK3006, GE-9 or HEP-2.





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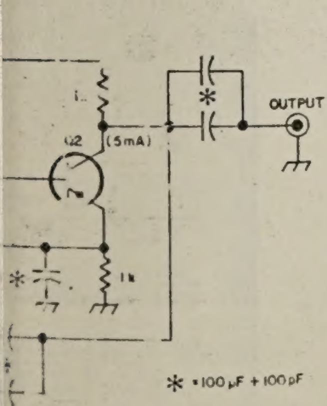
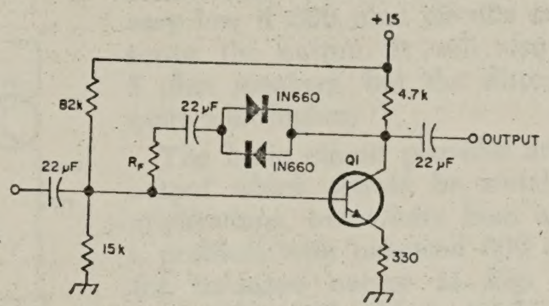
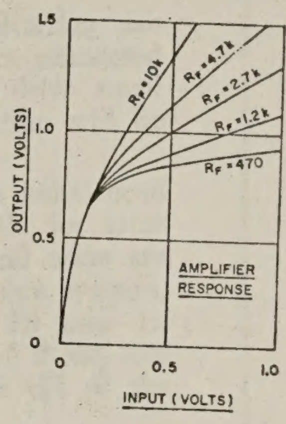


Fig. 6. The gain of this amplifier is controlled by the nonlinear feedback provided by two back to back diodes and the value of the feedback resistor R_f. Q1 is a 2N706, 2N708, 2N3394 or HEP-50.





Gain controlled amplifiers

It is a well known fact that the gain characteristics of an amplifier may be shaped by applying nonlinear feedback. In the amplifier of Fig. 6, the nonlinear feedback is furnished by two back to back diodes in the collector to base feedback path. Whenever the signal at the collector is high enough to forward bias the diodes (greater than approximately 0.6 volts peak to peak), negative feedback occurs and the gain of the amplifier is reduced. The gain of the stage may be further controlled by the value of the feedback resistor (R_f) as shown in the amplifier response curve. If it is desirable to have the nonlinearity occur at a higher level (greater than 0.6 volts peak to peak), more than one diode may be added to each leg of the feedback network. For lower levels, germanium diodes may be substituted for the silicon diodes specified in the schematic. With the germanium diodes in the feedback path, nonlinearity will occur when the signal is greater than about 0.1 volts peak to peak.

A voltage controlled, variable gain amplifier has many applications in automatic volume control, amplitude modulation and remote gain adjustment circuits. The only difference between the circuit shown in Fig. 7 and a standard common emitter amplifier is that a 1N34A diode is used in place of the emitter resistor. In an amplifier of this type, the gain of the stage is critically dependent upon the impedance of the emitter circuit. Since the impedance of the diode varies with the amount of current through it, the gain of the stage depends upon the transistor emitter current. The 1N34A was chosen because it provides an extremely wide impedance variation with a relatively gradual rate of change. This diode typically exhibits an impedance range from 15000 ohms at low levels to 200 ohms or less with high currents. The slow rate of impedance change is required to minimize distortion. This circuit is useful in ALC and AGC circuits, feedback regulation and other cases where wide dynamic range and instant response are required.

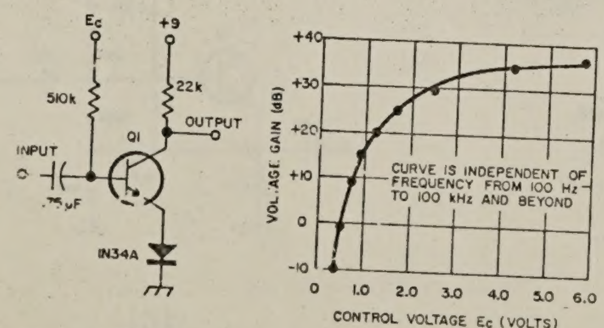


Fig. 7. Voltage controlled amplifier uses the varying impedance of a germanium diode in the emitter circuit to control gain. Transistor Q1 may be a 2N696, 2N3564, SK3019, GE-10 or HEP-54.

Preamplifiers

The simple low cost preamplifier in Fig. 8 provides extremely flat response from 10

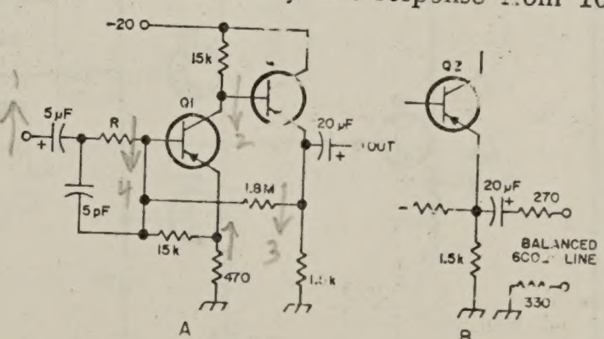


Fig. 8. High impedance preamplifier provides up to 1.2 megohms input impedance; the exact value depends upon the bleed-out resistor R. Both Q1 and Q2 should be a 2N2613, 2N2614, 2N2953, SK3004, GE-2 or HEP-254. A balanced output for reduced hum and noise may be obtained by using the padded output in B.

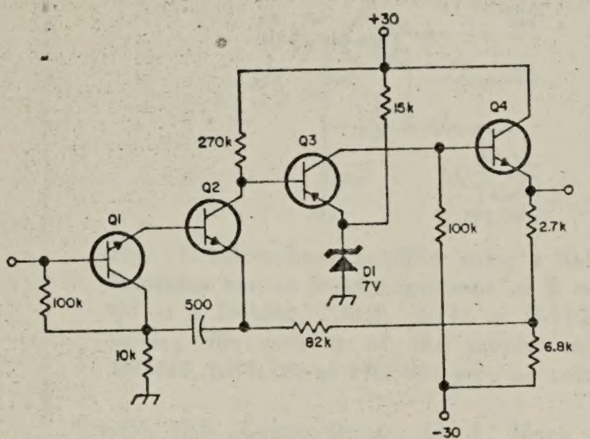


Fig. 9. This preamplifier provides 11 dB gain from 0.5 Hz to 2 MHz and has an input impedance of 32 megohms. Transistors Q1, Q2 and Q4 are 2N338, SK3020, or HEP-53; Q3 is a 2N328, GE-2 or HEP-52.

Hz to 30 kHz and at the same time exhibits an input impedance up to 1.2 megohms. The input impedance of the first transistor with the unbypassed emitter resistor is on the order of 50,000 ohms; by including the build-out resistor R in the circuit, the input impedance may be increased up to 1.2 megohms. Without R in the circuit, the voltage gain is approximately 15. As the value of R is increased, the voltage gain decrea.es and the entire circuit exhibits unity gain when the value of R is 1.2 megohms.

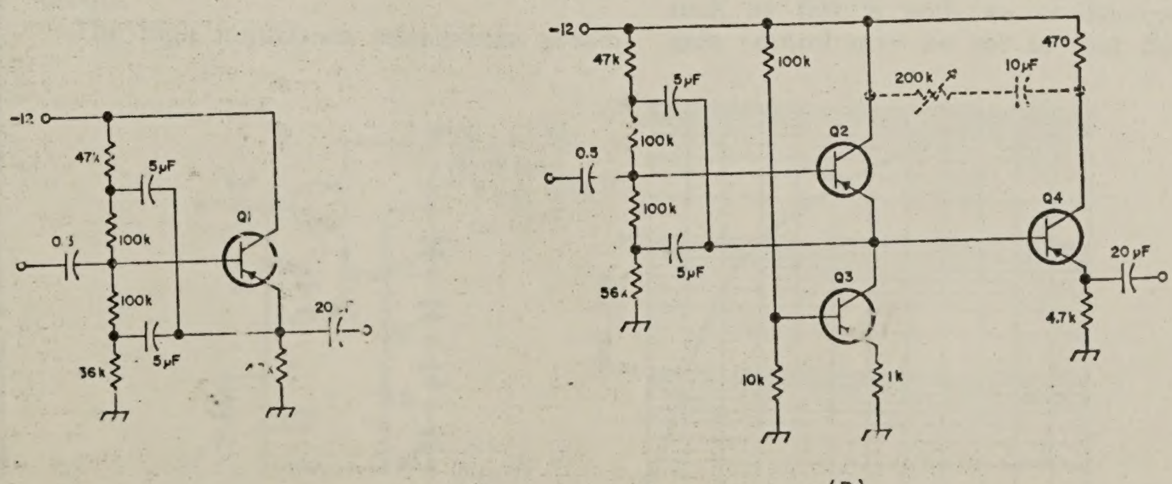
The output impedance of this simple preamplifier is particularly low, so it may be used for driving all types of circuits. Harmonic and intermodulation distortion are very low if 600 ohm circuits are connected across the output. It will also drive small 8 ohm speakers, but the distortion will be quite a bit higher.

The basic circuit provides an unbalanced output which should be suitable for most applications, but where hum and noise are a problem with balanced 600 ohm systems, the balanced output of Fig. 8B may be used. This pad adds a total of 6 dB loss in the output, but it does get rid of the hum and noise.

The four transistor preamplifier illustrated in Fig. 9 exhibits an input impedance of 32 megohms and provides 11 dB gain from 0.5 Hz to 2 kHz. The high input impedance of this amplifier is a function of the two negative feedback loops; one from the emitter of Q2 to the collector of Q1, the other from the junction of the 2.7k and 6.8k resistors in the emitter of Q4 to the emitter of Q2. The output impedance of this amplifier is 20 ohms so it may be used for driving many types of circuits.

In many cases an amplifier with an input impedance approaching that of a VTVM is required to keep circuit loading to a minimum. The amplifier of Fig. 10 more than meets these requirements; it provides up to 20 megohms input impedance, develops 1 volt rms across a 3300 ohm load and exhibits a frequency response from 10 Hz to 200 kHz.

The development of this circuit started



(A) (B)

.. 10. This high impedance preamplifier provides up to 20 megohms input impedance and has a frequency response from 10 Hz to 200 kHz. Circuit B was developed from circuit A by replacing the emitter resistor in A with Q3 and adding an emitter follower to reduce loading. The input impedance is further increased by the components shown by the dashed line. All transistors are 2N2188, SK3005, GE-9 or HEP-2.

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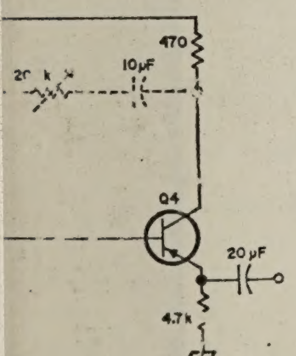

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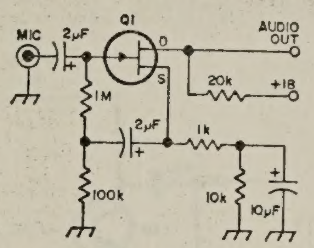


Fig. 11. Microphone amplifier using a field effect transistor has an input impedance of 5 megohms. Q1 is a 2N4360, TIM12, U-112 or U-110. By reversing the polarity of the supply voltage, a 2N3820, MPF-104 or HEP-801 may be used.

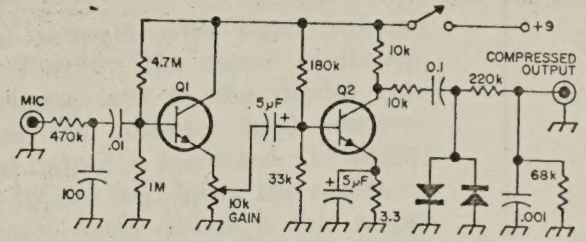


Fig. 12. Two stage clipper/preamp will increase the talk power of your rig. Transistors Q1 and Q2 are 2N1304, 2N2926, 2N3391, SK3011, or HEP 54. The diodes are 1N456 or HEP-158.

with the circuit shown in A. Here conventional bootstrapping was used on a basic emitter follower circuit to elimina' the shunting effect of the base bias resistors. When a transistor with a current gain of 100 was used, the input impedance was measured at 200K with a 3300 ohm load. A significant increase in input impedance may be obtained by replacing the emitter resistor of Q1 with the collector resistance of Q2 as shown in B. To keep the loading as light as possible on the emitter of Q1, an emitter follower (Q3) is used. With this circuit, the input impedance is slightly over 1 megohm with a 3300 ohm load.

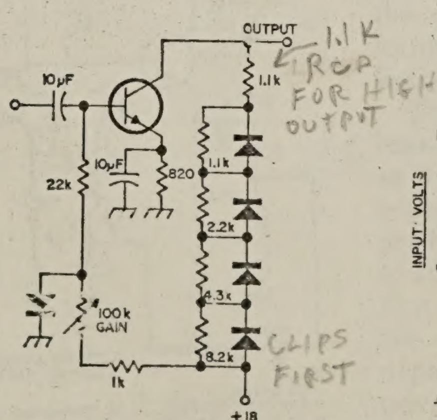
The input impedance of this circuit may be further increased with the addition of the components shown by the dashed lines. However, if this positive feedback is overdone, the circuit will oscillate. If, on the other hand, the 200k feedback pot is carefully adjusted, the input impedance may be raised to 20 megohms or so before instability occurs.

The high impedance microphone preamplifier illustrated in Fig. 11 makes use of the inherently high input impedance of field effect transistors. This impedance is raised still higher by the use of the 2 µF bootstrap capacitor from source to gate; in this case to about 5 megohms. This circuit's output impedance of 2k is suitable for driving other FET's or conventional junction transistors.

Clipper/preamplifier

The microphone clipper/preamp shown in Fig. 12 is very simple to construct and allows you to stay as far away from the mike as you like; it does a very good job of beefing up weak audio signals. It was designed primarily for high impedance dynamic microphones, but may be used with other mikes with slightly less gain. It provides up to 10 dB gain on low level audio signals and since it uses a minimum of parts, may be easily constructed in a small minibox.

Although the best way to adjust a clipper such as this is with an oscilloscope the gain control may be set so that the final



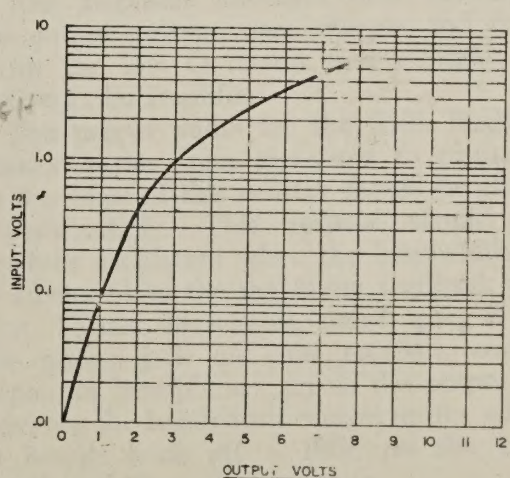


Fig. 13. This simple dynamic range compressor provides more than a 50 dB range; it exhibits gain with a 20 millivolt signal but will not saturate with input voltages up to 6 or 7 volts. All the diodes are 1N914, transistor Q1 should be a 2N2926, 2N3391, SK3010, GE-8 or HEP-54.

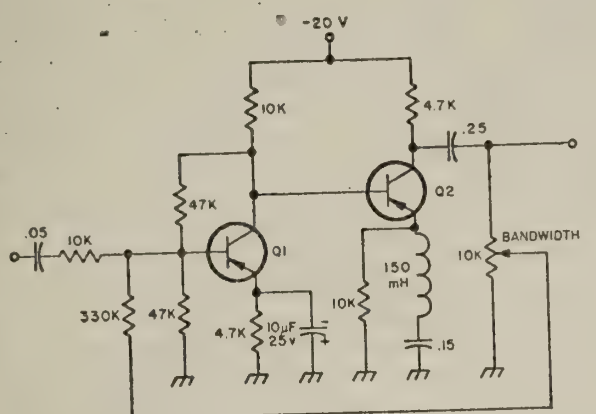


Fig. 14. This simple audio bandpass filter may be narrowed to the limits of unintelligibility. At a bandwidth of 80 Hz, it provides about 20 dB gain. The input is connected to the phone jack on your receiver while headphones are connected across the output.

audio stage (of your transmitter) approaches saturation on a steady whistle (into a dummy load please); this will approach optimum adjustment. A final check should be an on the air report from a nearby station so you can determine the approximate range settings appropriate for your particular transmitte.

Compression amplifier

The compession amplifier illustrated in **Fig. 13** provides a minimum output signal with only 20 millivolts (0.02 volts) input,

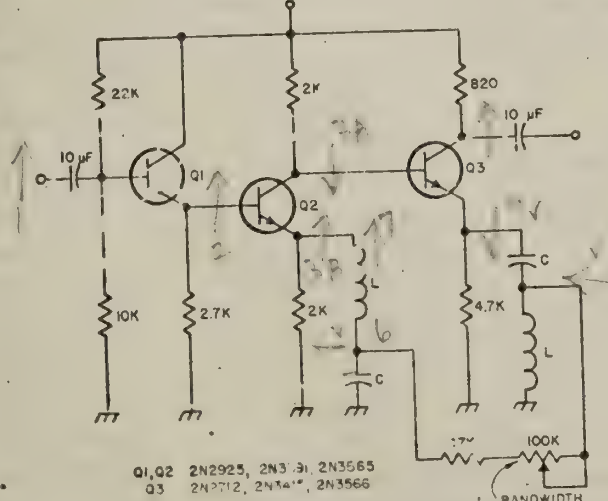


Fig. 15. This three stage audio filter uses two series resonant circuits to provide a very narrow audio passband. The Q of the circuits, and therefore the bandwidth, is controlled by the amount of feedback.

but will not saturate with input voltages up to 6 or 7 volts. The secret to its operation of course lays in the diodes connected across the collector load resistors. As the signal output is increased, the diodes conduct one by one and lower the resistance of the collector load. Although this amplifier has a minimum gain of 1 and a maximum gain of 15 with the components shown in the schematic, the gain characteristics may be made to follow other curves by the proper selection of load resistors and diodes.

Audio filters

The two transistor audio filter in Fig. 14 uses positive feedback to increase the Q of an inexpensive LC circuit to a very high degree. At a bandwidth of 80 Hz for example, this circuit provides 20 dB gain and furthermore, the bandwidth may be decreased to the limit of intelligibility. The gain stability is increased in this amplifier by the use of negative feedback from collector to base of Q1; this also serves to reduce the output impedance and increase the power transfer to the succeeding stage.

At frequencies far removed from the resonant LC circuit in the emitter of Q2, the emitter impedance is essentially that of the 10k emitter resistor. As resonance however, the low series impedance of the LC network predominates and increases the gain of the stage. Since the output signal is in phase with the input signal, the feedback through the 10k bandwidth pot and 330k resistor is regenerative. As the gain of the amplifier increases near resonance, the output voltage rises sharply and transforms the low Q circuit into a highly selective audio amplifier.

The proper value for the 330k feedback resistor varies from transistor to transistor, so the value of this resistor should be chosen experimentally. This resistor should *just* produce oscillation when the bandwidth pot is advanced to the maximum feedback position. To use this circuit, simply plug it into the phone jack on your receiver, connect a pair of headphones across the output and advance the bandwidth control until a whistle is heard; back off a little on the bandwidth and it's ready to operate.

The audio filter illustrated in Fig. 15 is somewhat similar to the one in Fig. 14 except that the passband has a better shape

Fig. 16. This highly ve
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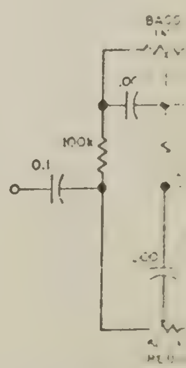

Fig. 17. Resistance cap
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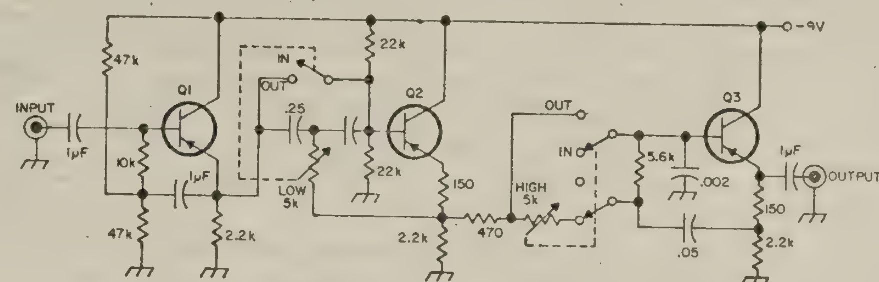
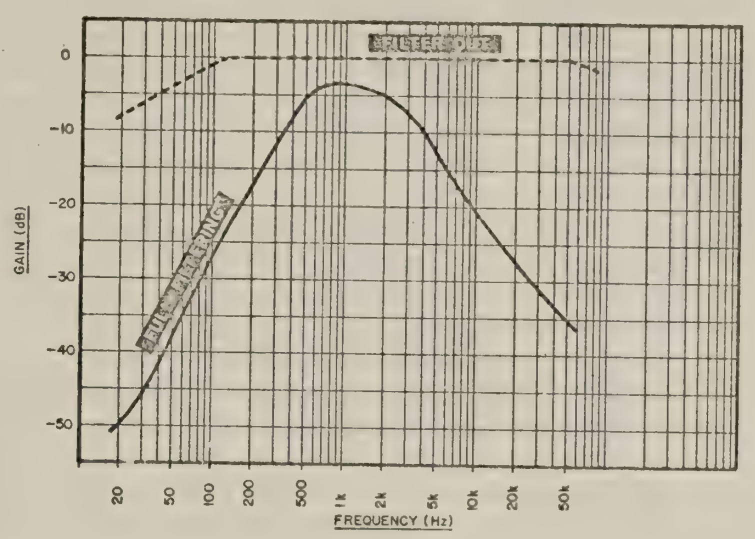


Fig. 16. This highly versatile audio filter may be used to completely shape the audio spectrum of your receiver or transmitter; it may be used with the filters out, or with the variable low- and high-pass filter networks connected. All the transistors are low cost audio types such as the 2N1305, 2N1380, 2N2613, GE-2 or HEP-253.



factor because two series resonant LC circuits are used. Here again feedback is used to raise the Q of the resonant circuits to a very high value. The 47k resistor in the feedback line should be adjusted experimentally so that the circuit will just oscillate when the 100k bandwidth pot is shorted out (maximum clockwise position). Although the LC values shown in the schematic are for a center frequency of approximately 1000 Hz, other values may be used for

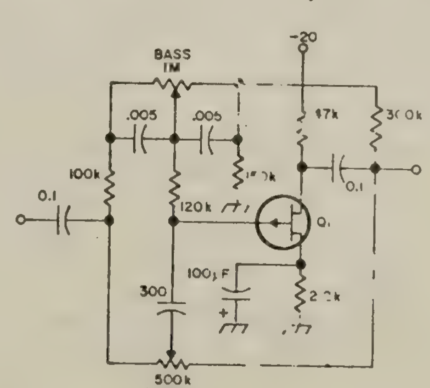


Fig. 17. Resistance-capacitance tone controls are usually not too satisfactory with junction transistors because of heavy loading. The high impedance characteristics of the FET eliminates this problem with no loss in the dynamic range of the tone control. Q1 is a 2N2943, 2N3820, MPF-105 or HEP-801.

other center frequencies.

One of the problems in amateur SSB communications is that the audio spectrum of the speech amplifier should be shaped so that it amplifies only those signals between about 300 and 3000 Hz. This can be accomplished by high-Q tuned circuits, but the inductors required are quite large and expensive. A simpler approach is to use the adjustable audio bandpass filter shown in Fig. 16. When the high- and low-pass filter of this amplifier are out of the circuit, it is flat within 1 dB from 100 Hz to 50 kHz. With the filters in the circuit, the audio may be shaped between the limits shown in Fig. 16.

Tone control

Audio tone controls using conventional junction transistors are difficult to build because the low input impedance of these devices seriously limits the tone control's dynamic range. An obvious solution to this problem lies in applying the inherently high input impedance of the field effect transistor. The tone control illustrated in Fig. 17 should be familiar to old vacuum tube hands: it is a straightforward tone control for both treble and bass using a modern FET in place of a thermionic triode.

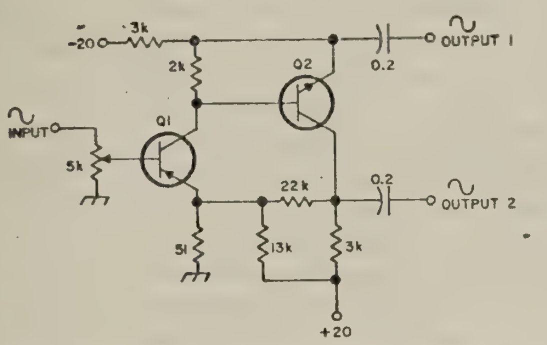


Fig. 18. This phase splitting circuit provides two out of phase signals for driving a push pull amplifier without an expensive transformer. The gain of the stage as shown is 150, but this may be adjusted by changing the value of the 22K feedback resistor. Q1 and Q2 are a complimentary pair such as the 2N652 and 2N388 or 2N2430 and 2N2706.

Phase splitter

The simple phase splitting circuit in Fig. 18 is a two stage direct coupled amplifier connected as a complementary pair with feedback and illustrates a novel way of obtaining out of phase driving signals for a push pull amplifier without an expensive transformer. The input transistor is a common emitter voltage amplifier with its collector tied directly to the base of Q2. The 3k resistor in the emitter of Q2 provides bias for this transistor but does not cause regeneration because it is common to both the base *and* emitter. The 13k resistor sets the overall circuit bias and its value is chosen so that the collector and emitter of Q2 are at the desired operating level. The 22k feedback resistor provides negative feedback to the emitter of Q1 and determines the gain of the circuit. In this case 22 kilohms was chosen to set the gain at 150, but other values of gain may be obtained by adjusting the value of this resistor.

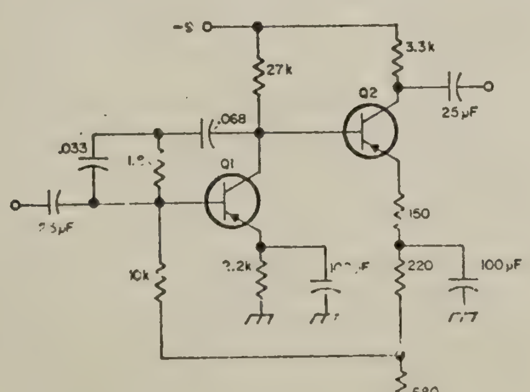


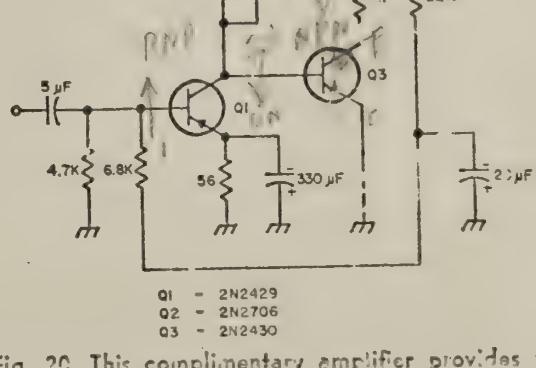
Fig. 19. This phono preamplifier uses frequency selective feedback between the collector and base of Q1 to obtain proper equalization during playback. Transistors Q1 and Q2 are 2N1384, SK3003, GE-2 or HEP-254.

Equalized audio amplifier

The equalized audio amplifier shown in Fig. 19 is a two stage direct coupled audio amplifier with a frequency selective feedback path. It is particularly suitable for boosting and equalizing the signal from a ceramic phono pickup to obtain a *flat* output of sufficient level to drive an audio power amplifier.

When playing a record, the output from the pickup is proportional to the force to which the stylus is subjected when tracing the groove. In fact, the open circuit voltage across the pickup is approximately proportional to the logarithm of the frequency with reference to the recorded amplitude. If the pickup is loaded with a very high impedance on the order of one or two megohms, the output versus input is nearly the inverse of the recording characteristic; therefore, the equalization is automatic.

However, it is not always possible to load the pickup with a very high impedance circuit, especially when transistors are used



Q1 - 2N2429
Q2 - 2N2706
Q3 - 2N2430

Fig. 20. This complimentary amplifier provides up to 220 mW output with a frequency response from 90 Hz to 12.5 kHz. Although matched transistors are not required for Q2 and Q3, they are available as the 2N2707.

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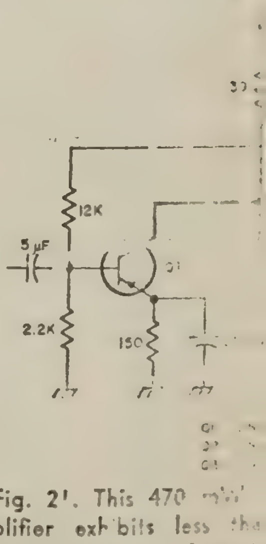
In the amplifier [...] is obtained by a [...] back path between [...] of Q1. Stabilizatio[...] the direct coupling [...] the current feedba[...] ohm resistor in th[...] tion, more negati[...] by the unbypassed [...] second stage.

Complementary [...]

The small transf[...] amplifier illustrated [...] output of 220 mW [...] 40 microamps. The [...] ended class B outpu[...] common collector [...] biased by a resisti[...] the driver transist[...] resistors in the out[...] quate temperature [...] lished by cut and [...] ohms seems to off[...] In adjusting this a[...] pot should be adju[...] current of the ou[...] order of 2.5 mA; [...] mum of crossover d[...] adjusted, this ampli[...]



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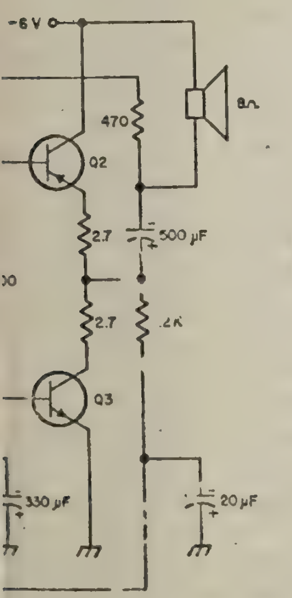

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in the preamplifier. Resistive pads may be used to increase the input impedance, but they greatly reduce stage gain and increase problems with signal to noise ratios.

In the amplifier of Fig. 19 equalization is obtained by a frequency selective feedback path between the collector and base of Q1. Stabilization at dc is provided by the direct coupling between Q1 and Q2 and the current feedback path through the 10 ohm resistor in the base of Q1. In addition, more negative feedback is provided by the unbypassed emitter resistor in the second stage.

Complementary power amplifiers

The small transformerless complementary amplifier illustrated in Fig. 20 provides an output of 220 mW with an input of only 40 microamps. The transistors in the single ended class B output stage are used in the common collector configuration and are biased by a resistive voltage divider and the driver transistor circuit. The emitter resistors in the output stage provide adequate temperature stability and are established by cut and try, but a value of 2.7 ohms seems to offer a good compromise. In adjusting this amplifier, the 100 ohm pot should be adjusted so that the idling current of the output transistor is on the order of 2.5 mA; this will insure a minimum of crossover distortion. When properly adjusted, this amplifier will exhibit a ±3 dB frequency response from 90 Hz to 12.5 kHz and distortion of 4% at 120 mW input and 10% at 220 mW input.

Another complementary audio amplifier is shown in Fig. 21. This power stage provides 470 mW output and utilizes both ac and dc feedback to minimize distortion and extend frequency response. Although matched transistors are not required for the proper operation of this amplifier, a set of matched transistors is available as the 2N2707; the cost of the matched pair is only several cents more than the total separate costs of a 2N2706 and 2N2430. This very useful amplifier is flat ±3 dB from 15 Hz to 130 kHz, exhibits an input impedance of 750 ohms and produces less than 2% distortion at 470 mW output.

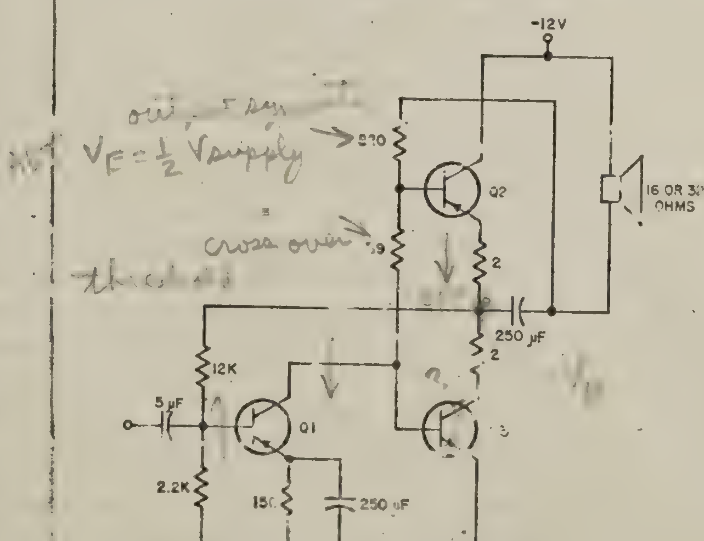


Fig. 21. This 470 mW complementary audio amplifier exhibits less than 2% distortion and is flat within ±3 dB from 15 Hz to 130 kHz.

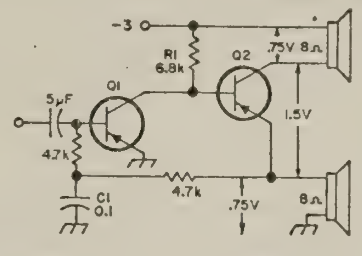


Fig. 22. An audio power amplifier with push pull output using a single transistor in the final stage may be obtained with this simple circuit. Only about 50 mW is available from this amplifier, but the gain is flat up to 30 kHz. Both Q1 and Q2 should be germanium audio transistors such as the 2N215, 2N404, 2N2953, SK3004 or HEP-253.

Single transistor push pull

The amplifiers illustrated in Figs. 22 and 23 illustrate how a quasi-push-pull output may be taken from a single transistor.

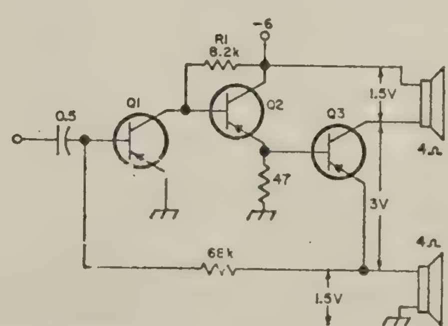


Fig. 23. A larger power version of the single transistor push pull circuit of Fig. 22 is shown here. The operating characteristics are similar to the 50mW circuit except that approximately 1 watt may be obtained. Transistors Q1 and Q2 are 2N215, 2N404, 2N2953, SK3004 or HEP-253; Q3 is a 2N554, 2N1032, 2N1666, SK3009 or HEP-232.

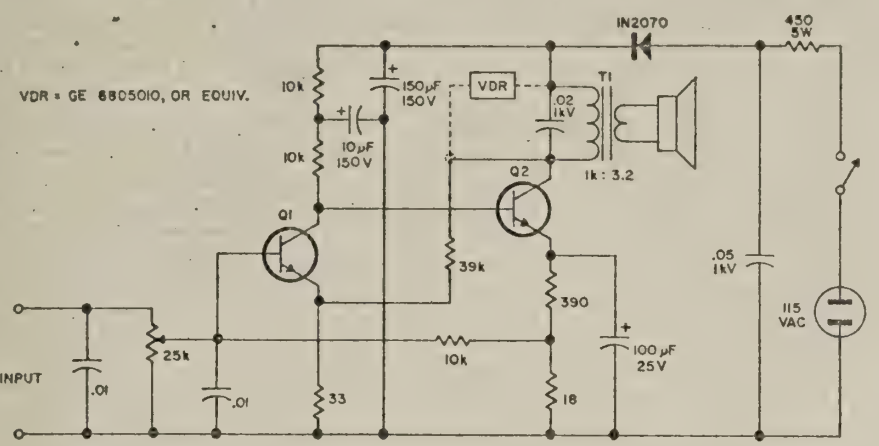


Fig. 24. This line operated audio power amplifier provides about 500 mW output with an 80 millivolt input signal. Q1 is a 2N3565, SE4002, SK3020 or HEP-54; Q2 is a 2N3916 or SE7005.

These amplifiers are dc connected, thereby eliminating many components, while at the same time assuring excellent low frequency response.

In the circuit in Fig. 22, the first transistor serves as both an ac driver and part of the dc bias system. Although the values shown in the schematic were selected for optimum results, the 6800 ohm biasing resistor (R1) should be adjusted experimentally to obtain equal voltages across the collector and emitter loads as illustrated in the schematic.

The frequency response of this amplifier may be adjusted by changing the value of the 0.05 μF capacitor (C1). When this capacitor is left completely out of the circuit, the heavy negative feedback around the circuit provides a frequency response that is flat from dc (with the input capacitor shorted) to 30 kHz. However, under these conditions, the gain is only about 35 dB. The maximum power output available from this circuit is on the order of 50 mW; above this level severe clipping occurs with noticeable audible distortion.

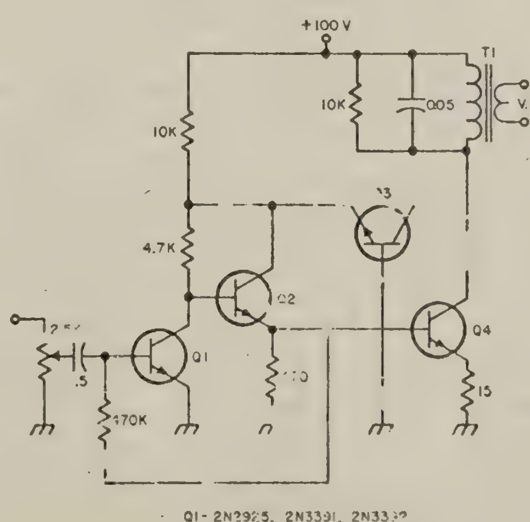
A higher power circuit that exhibits essentially the same characteristics is illustrated in Fig. 23. With properly heat sinked transistors, this unit provides usable outputs up to one watt. As with the lower power circuit, the bias resistor (R1) should be adjusted to provide equal voltages across the emitter and collector loads shown in the schematic.

The input impedance of both these circuits is on the order of several thousand ohms, so they may be easily driven by other transistor circuits. Although two separate speakers are shown as the output load, the load could just as easily be two separate transformer windings.

Line operated amplifiers

The line operated one watt amplifier in Fig. 24 provides about 500 mW output with an 80 mV input signal. The use of transistors with high collector to emitter voltage ratings permits the use of a transformerless power supply operating directly from the 115 volt ac line. To prevent damage to Q2 in the event of transient voltage spikes on the line, a voltage dependent resistor (VDR) such as a General Electric Thyrector or Motorola Thyristor should be connected across the primary of the output transformer.

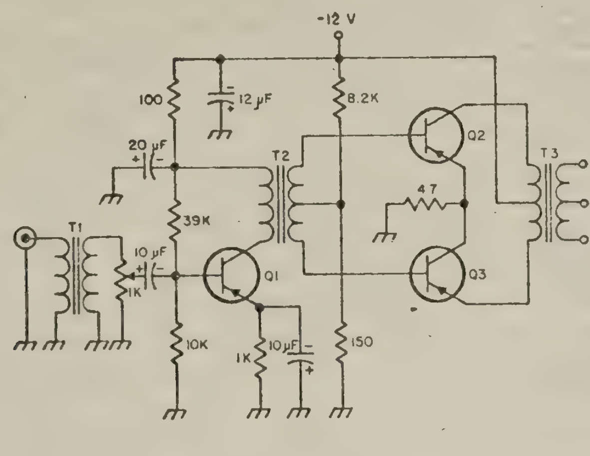
Another line operated power amplifier is illustrated in Fig. 25. This amplifier is



Q1- 2N2925, 2N3391, 2N3392
Q2- 2N2924, 2N2712, 2N3393
Q3- 2N2923, 2N2926
Q4- 2N4054
T1 -2500 ohms TO VOICE COIL

Fig. 25. High performance with low cost is obtained with this line operated audio power amplifier because expensive electrolytic capacitors are eliminated by direct coupling between stages. This circuit delivers 1 watt to the speaker with 3 mW input.

ters. The circuit is relatively straight forward, with a single audio amplifier driving the class B push pull power stage through a small transformer. To modulate the collector of a small transmitter, simply run the collector voltage supply through the secondary of the "moduation" transformer, in this case a low cost 5k:200 ohm audio transformer.



Q1,Q2,Q3 2N109, 2N2613, 2N2953, SK3004
T1 25K:1200 OHMS
T2 20K:2K OHMS CT
T3 5K CT:200 OHMS CT

Fig. 26. This 100 mW modulator may be used to collector modulate transmitters up to about 200 mW or to base modulate somewhat larger power amplifiers. Good performance with a minimum of components is obtained by transformer coupling between stages.

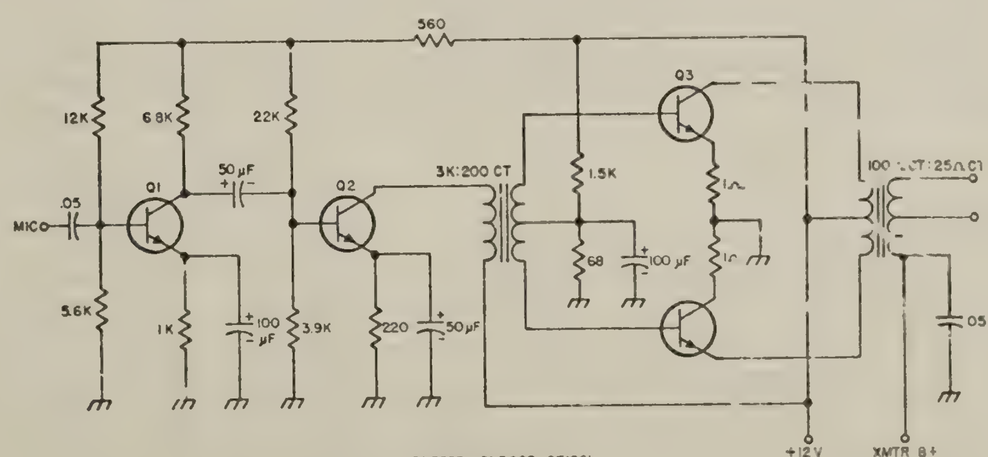
based on the use of a high voltage plastic transistor, the 2N4054. The circuit delivers one watt of audio power to a speaker with about 3 millivolts input signal; at this power level the total harmonic distortion at 1 kHz is less than 10%. The key to its low cost performance is the fact that direct coupling is used, thereby eliminating the need for expensive electrolytic capacitors.

Modulators

The 100 milliwatt modulator illustrated in Fig. 26 is suitable for collector modulating small transistor transmitters up to about 200 milliwatts. It may also be used for base modulating somewhat larger transmitters.

The 5 watt modulator shown in Fig. 27 may be used to modulate transmitters with up to 10 watts input. The use of low cost, high gain silicon transistors and efficient transformer coupling significantly decreases the complexity of the circuit. Usually many more transistors are required to obtain five watts of audio with a microphone input. Although this modulator was designed for a ceramic or crystal microphone, it may be used with dynamic types with slightly less gain. This circuit exhibits extremely low distortion characteristics, and when used to collector modulate a ten watt transistor transmitter, produces extremely clear and crisp audio.

The transistorized 25 watt modulator shown in Fig. 28 is not much different from other types which have been described, but with three transformers it is somewhat more efficient than most. The transformers are readily available commercial models which may be obtained from most suppliers. However, transformer T2 must have a center tap on the secondary; this is easily accomplished by unwinding 46 turns from the outside winding, bringing out a center tap at this point and rewinding. Impedance matching to the rf amplifier is accomplished by adjusting the rf output loading network.



Q1 2N3393, 2N3693, SE1001
Q2 2N3404, SE6001
Q3,Q4 2N2868, 2N3110, 2N3523

Fig. 27. 5 watt modulator for transmitters up to 10 watts input. High gain silicon transistors and transformer coupling increase performance at decreased circuit complexity.

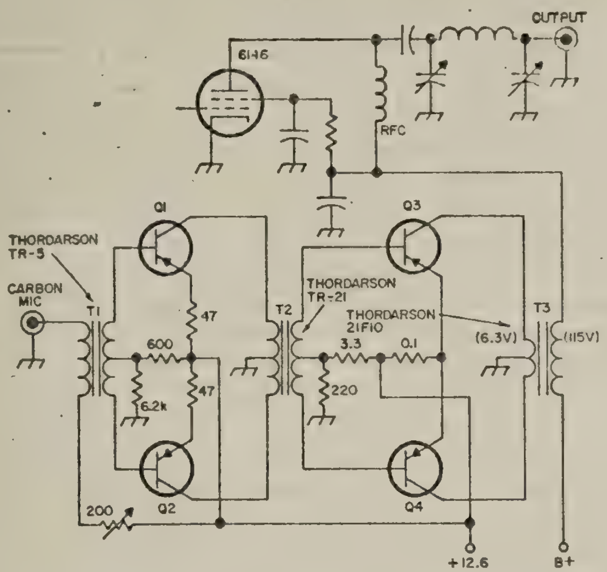


Fig. 28. 25 watt modulator uses readily available commercial transformers. Transistors Q1 and Q2 are 2N1172, 2N301, 2N1560, SK3009, GE-9 or HEP-232; Q3 and Q4 are 2N174, 2N278, SK3012, GE-4 or HEP-233.

The frequency response of this circuit is quite good and is essentially flat from 200 Hz to 7 kHz.

Circuits for Receivers

Injection oscillator

If you are interested in building a dual conversion receiver for single channel operation on MARS, 2 meter FM or WWV, the single oscillator circuit which provides two outputs illustrated in Fig. 29 should be of interest. This single oscillator circuit results in a reduction of components without sacrificing receiver performance. Basically it consists of a conventional common base transistor oscillator which provides the injection voltage for the second mixer. The output of the oscillator is fed into a diode harmonic generator and resonant tank which is tuned to the desired harmonic; this harmonic is used for injection into the first mixer. For example, for a dual conversion 15 MHz WWV receiver with a 455 kHz if, a 3636 kHz crystal would be used along with its third harmonic at 10.908 MHz. The 10.908 MHz signal would be mixed with the 15 MHz WWV signal in the first mixer to provide an output at 4092 MHz; this signal would in turn be mixed with the 3636 kHz oscillator output in the second mixer to provide the 455 kHz *if*. The only consideration in choosing the crystal and harmonic frequencies is that only *odd* harmonics should be used. This is because when even harmonics are used in this scheme, poor second *if* image rejection will be a problem.

In most receivers the oscillator injection frequencies are below that of the signal frequencies. This is usually desirable since it results in a lower first *if* frequency which will provide better image rejection. In this case the necessary crystal frequency may be found from the following formula:

$$f_o = \frac{f_s - f_{if}}{h + 1}$$

Where: f_o = Crystal frequency

f_s = Signal frequency

f_{if} = *if* frequency

h = Harmonic to which diode tank circuit is tuned.

In those cases where it might be desirable to have the injection frequencies higher than the signal frequencies, the following formula may be used:

$$f_o = \frac{f_s - f_{if}}{h - 1}$$

BFO's

The simple BFO in Fig. 30 may be added to an existing receiver with a minimum of cost and effort. Essentially it is a tuned collector oscillator with an *if* transformer being used for the tuned circuit inductance. Just pick a transformer that is compatible with the *if* in your receiver; it doesn't make any difference to the transistor. Anything between 85 kHz and 1600 kHz will work

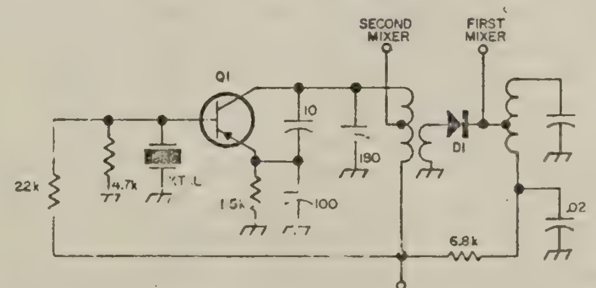


Fig. 29. Single oscillator and diode provide two injection frequencies for dual conversion receivers. Transistor Q1 is a 2N1745, 2N2188, TI410, GE-9 or HEP-2; the diode should be a 1N32A or similar.

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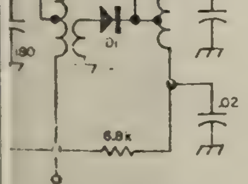


diode provide two
conversion receivers.
N2188, TIM10, GE-9
a 1N82A or si-ilar

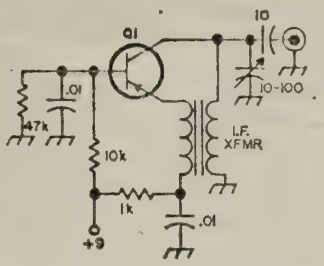
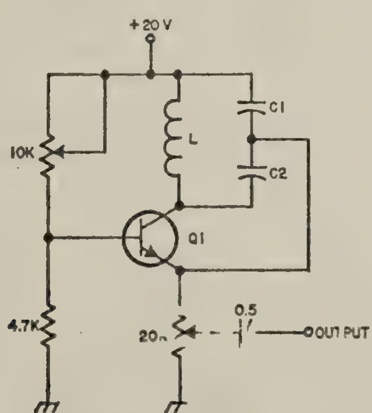


Fig. 30. This beat frequency oscillator may be added to existing receivers with a minimum of difficulty. The BFO frequency is determined by the *if* transformer which provides feedback from collector to emitter. Transistor Q1 should be a 2N384, 2N1749, 2N2362, TIM10, SK3008, GE-9 or HEP-2.

well in this circuit. Before you can use the transformer though, remove all of the fixed tuning capacitors from the unit; usually these are readily available on the bottom of the transformer. If the circuit does not oscillate when voltage is applied, reverse the transformer leads going to the emitter of the transistor. To connect the BFO into the receiver, run a piece of small coaxial cable from the BFO output to the base (or grid) of the detector. In some cases sufficient injection will be obtained by just placing the coax lead in the immediate vicinity of the detector. Adjust the core in the BFO transformer so that the variable tuning capacitor allows the BFO output to swing to either side of the receiver *if*; then the variable capacitor will operate as a pitch control.



Q1 - 2N2925, 2N3392

FREQUENCY	C1	C2	L
50 kHz	3500 pf	1500 pf	10 mH
80 kHz	2200 pf	910 pf	6.2 mH
100 kHz	1800 pf	750 pf	4.7 mH
200 kHz	910 pf	390 pf	2.2 mH
455 kHz	390 pf	160 pf	1 mH
1000 kHz	180 pf	75 pf	0.47 mH

Fig. 31. This simple circuit provides an extremely stable BFO. The frequency of oscillation may be tailored to your needs by simply choosing the proper tank components listed in the table.

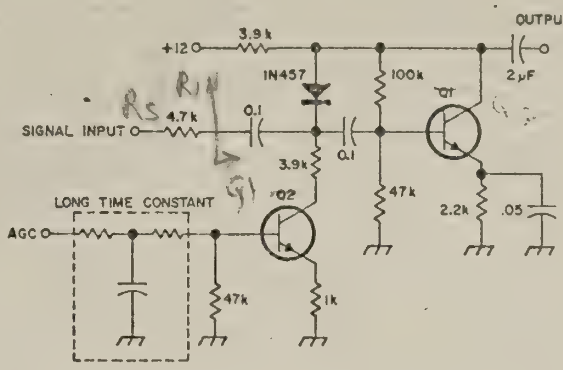


Fig. 32. This super AGC circuit only requires two transistors to obtain up to 60 dB of control. Q1 and Q2 are 2N1613 or HEP-254.

The circuit illustrated in Fig. 31 represents a temperature stable Colpitts oscillator which is very useful as a BFO. This oscillator utilizes an inexpensive silicon planar transistor and is exceptionally stable over wide ranges in temperature. In addition, it is characterized by a large output amplitude (10 volts peak to peak) and low harmonic distortion. In addition to duties as a beat frequency oscillator, this circuit is useful where a stable signal source is required up to several MHz. The 20k emitter pot is an output level control; the 10k pot in the base bias leg is used to adjust the base bias for maximum amplitude output.

AC neg FB from Q2 coll to Q2 base lowers Rin and causes drop across 4.7K Rs

AGC circuit

The super AGC circuit shown in Fig. 32 requires only two transistors to obtain 60 dB of control, while maintaining low distortion and power requirements. When a signal appears at the input with its corresponding AGC signal on the input of the long time constant circuit, Q1 conducts and causes current to flow through the diode. This current flow lowers the impedance of the diode and provides a low impedance path between the amplified signal at the collector of Q2 and the input signal at its base. When the diode conducts heavily, there is nearly 100% feedback and the gain of this stage is nearly unity. In addition, the input impedance of the stage becomes very low and results in a large voltage drop across the series connected 4700-ohm resistor.

When the collector current of Q1 is reduced to zero, the feedback through the diode is negligible and the gain and input impedance of Q2 is the same as that of a standard common emitter amplifier. With

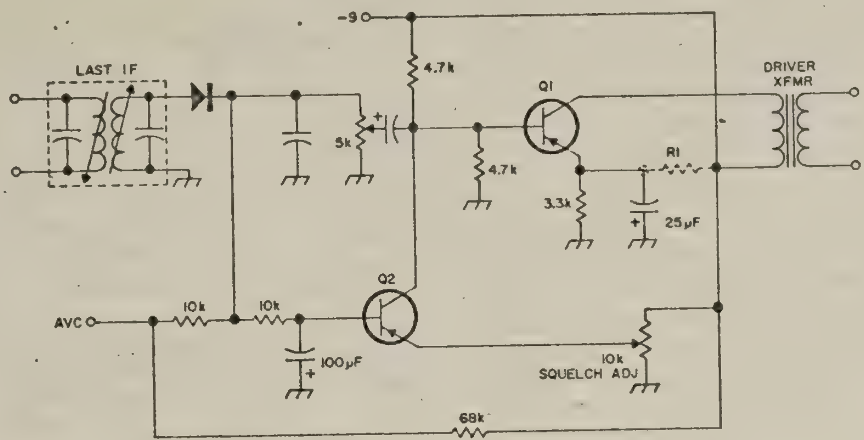


Fig. 33. This simple squelch circuit may be added to any transistorized receiver with only minor changes; in most receivers only four additional components are required, two resistors, a capacitor and a transistor. Q1 is an existing transistor in the receiver; Q2 is a 2N404, 2N2953, SK3004, GE-2 or HEP-254.

the constants shown in the schematic, a 60 dB control range is provided by a 0 to 100 µA AGC input. To increase this range, high gain transistors and a higher voltage power supply must be used.

Squelch circuits

The simple, but positively acting squelch circuit in Fig. 33 may be added to any transistorized receiver with only minor changes in the audio section and four additional components. Without an input signal, normal forward bias to the *if* amplifier flows in the AGC line. A portion of this bias voltage is applied to the base of Q2 through the 10K ohm squelch adjust. This voltage biases Q2 into full conduction with the squelch control pot determining the degree. When Q2 is saturated, base bias for Q1 is diverted to ground so the driver cannot amplify incoming noise and the speaker is quiet. When a carrier large enough to cut off Q2 is received, Q1 conducts and amplifies normally. The 100 µF filter capacitor in the base of Q2 removes all but the AGC signal coming from the detector.

To make the squelch less sensitive to large noise pulses, resistor R1 will ensure that transistor Q1 will be cut off until rf operates the squelch. The value of this resistor should be determined experimentally, since its value depends upon the type of transisto used in this stage.

Another simple squelch circuit is illustrated in Fig. 34. When there is no signal

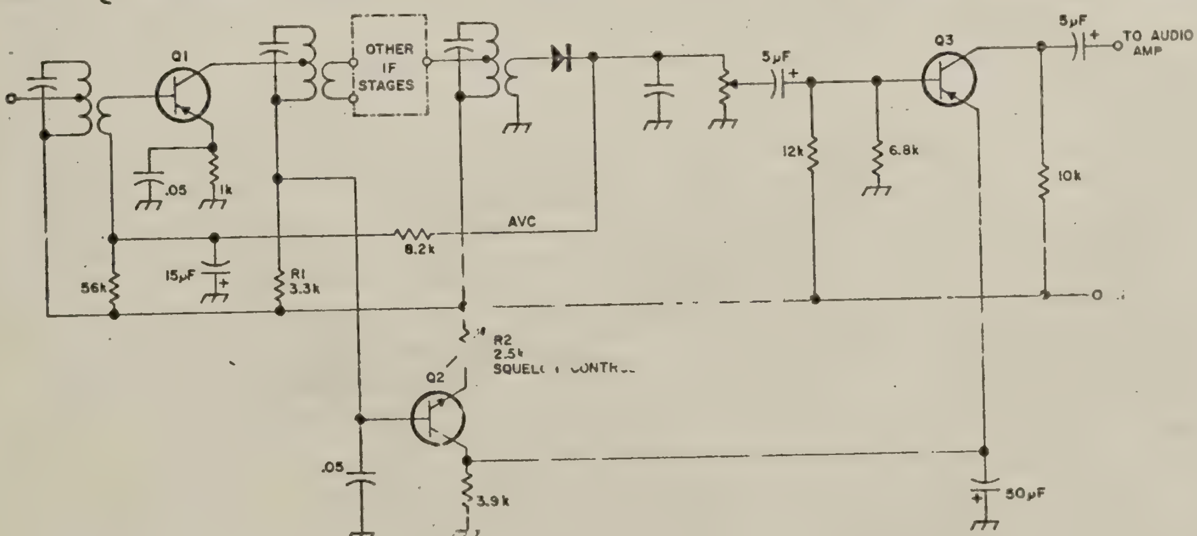


Fig. 34. This squelch circuit is very versatile and is capable of squelching out a 300 µV signal and still maintain control down to less than 1 µV. With the 2.5K pot set to squelch out a signal, approximately 3 dB increase in received signal will override the squelch. Q2 is a 2N1304, GE-5 or SK3011; Q3 is a 2N1274; GE-2 or HEP-254.

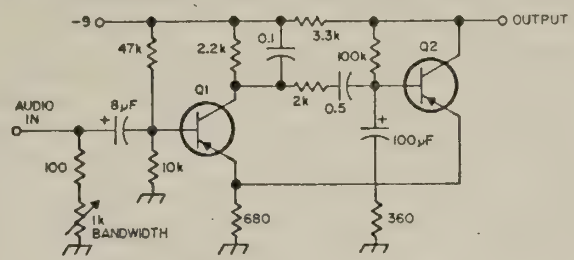


Fig. 35. This audio filter uses a 1000 Hz Wien bridge circuit to provide bandwidths from 70 to 600 Hz wide. Q1 and Q2 are 2N408, 2N2613, SK3004, GE-2 or HEP-254.

coming into the receiver, the AGC controlled *if* amplifier Q1 is operating at maximum gain. Under this condition the voltage drop across R1 is also maximum and is used to turn Q2 on. The current through Q2 is determined by the squelch control, R2. If R2 is set so that the current flow through Q3 causes the voltage drop across the 3.9k emitter resistor to be greater than the base voltage of Q3, it will be turned off. Under these conditions, there will be no output from Q3 and the receiver will be quiet.

When a signal is picked up by the receiver and the AGC is operating, the current through Q1 is reduced according to the signal strength of the received signal. This in turn reduces the voltage across R1, reduces the current through Q2 and lowers the voltage on the emitter of Q3. When the emitter to base junction of Q3 is forward biased, the stage turns on and results in an output to the audio amplifier.

This circuit is very versatile and is capable of squelching out a 300 μV signal into the receiver and still maintain control down to less than 1 μV. In some cases there may be some audio distortion between the on and off conditions of Q3, depending upon the setting of the audio gain control. This distortion is not ordinarily objectionable however and when the signal is several times greater than that required to just trigger the squelch, it is not present.

Selective audio amplifier

Selective transistor amplifiers are very helpful in sorting out stations from the QRM that plagues our HF bands. They are also quite helpful in VHF and UHF work for effectively narrowing the bandwidth of the receiver. This is because as the bandwidth is narrowed, the noise in the receiving bandpass is reduced accordingly.

In the selective transistor amplifier illustrated in Fig. 35, the frequency selected is determined by a modified Wien bridge circuit in the collector of the first transistor. Although the constants shown in this circuit are for a center frequency of 1000 Hz, other frequencies may be selected by the proper choice of bridge components. The bandwidth of this circuit is determined

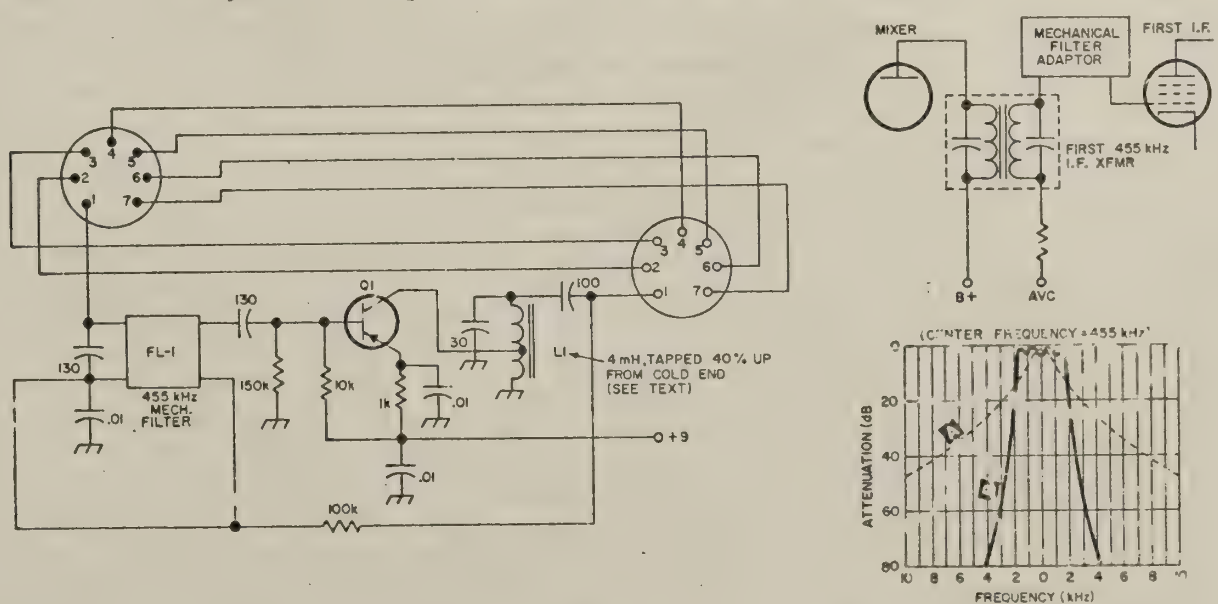


Fig. 36. The selectivity of inexpensive communications receivers may be substantially increased by the addition of this mechanical filter adapter. The transistor is used to make up for the 10 dB loss through the filter. The typical passband of a receiver without the filter is shown by A in the frequency response curve; the mechanical filter adapter results in curve B. Q1 should be a 2N1638, 2N1727, SK3003 or HEP-3.

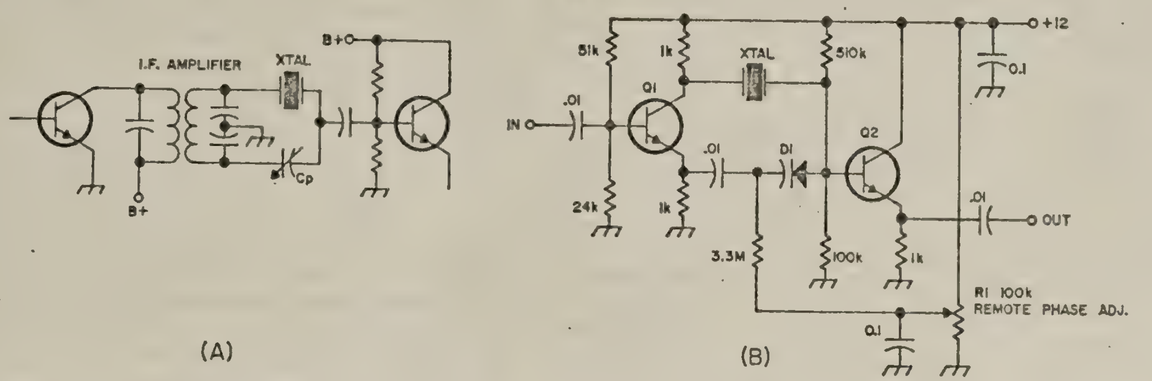


Fig. 37. Usually the crystal filter circuit in a receiver (A) must be physically located so the phasing capacitor (C_p) is accessible to the front panel. By using the varactor phased filter in B, the crystal may be located in any convenient location. Q1 and Q2 are 2N3478, 2N3564, 2N3707, 40236 or HEP-50; D1 is a 20 pF varactor such as the 1N954 or TRW V20.

by the value of the input load resistor, but with the resistors shown, it may be varied between 70 and 600 Hz.

Mechanical filter adapter

With the heavy QRM that is rampant on today's high frequency ham bands, the selectivity of many of the lower cost communications receivers leaves a great deal to be desired. In fact, in many cases the selectivity of these receivers is hopelessly inadequate. Adding a Q multiplier or a simple crystal filter will help to some extent, but these devices simply narrow the peak of the *if* response curve. Although this is quite suitable for CW work, it is of little help in separating SSB and AM stations. A much more useful improvement is the addition of a mechanical filter to the receiver *if*. Unlike simple LC circuits, the mechanical filter closely approximates the ideal bandpass response curve.

Wiring this filter into a receiver could require some pretty extensive rework, but by using the transistorized mechanical filter adapter illustated in Fig. 36, it may be simply plugged into the first *if* amplifier tube socket. The actual circuit itself is very straightforward; the simple transistor amplifier makes up for the 10 dB of loss through the mechanical filter. Coupling the output of the transistor to the grid of the first *if* tube is accomplished with the 4 mH coil. This may be made by winding 100 turns of number 36 on a cup core or toroid (tapped at 40 turns) or the primary of an inexpensive 455 kHz if transformer may be used.

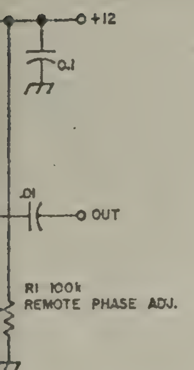
Layout of the circuit is not at all critical except that care should be taken to make sure that there is no leakage around the mechanical filter and amplifier. To this end the plate lead of the *if* tube should be shielded. Although the base layout shown in the schematic is for a 6BA6 tube, this adapter may be used with any *if* tube by simply placing the mechanical filter in the grid lead. Packaging this device is quite simple too; just mount an appropriate tube socket on top of a plug-in can (Vector G2.1-8-4), build the circuitry inside, plug it in the receiver tube socket and pick the weak signals out of the QRM.

Tunable crystal filter

One of the problems encountered when installing a crystal filter in a receiver for added selectivity is the fact that the unit must be installed physically close to the front panel so that the phasing control (C_p) is accessible. This problem is neatly solved by using the varactor tuned unit shown in Fig. 37. In this circuit, the crystal is phased by the varactor diode which may be remotely controlled by the variable resistor R1. The circuit may be used for any *if* from 100 kHz to 1500 kHz by simply selecting a crystal which is resonant at the desired frequency. In addition, the filter may be completely removed from the system by simply forward biasing the diode.

Cascode amplifiers

One of the big advantages of the cascode rf amplifier is that in the high frequency range it does not require neutralization. In the cascode circuit shown in Fig. 38, two transistors are connected so that the mismatch between them reduces internal



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2N3564, 2N3707,

kage around the ifier. To this end tube should be se layout shown 6BA6 tube, this any *if* tube by ical filter in the device is quite appropriate tube can (Vector G2.1- nside, plug it in nd pick the weak

ncountered when n a receiver for ict that the unit lly close to the asing control (C_p) is neatly solved ed unit shown in crystal is phased hich may be re- variable resistor used for any *if* Hz by simply se- resonant at the dition, the filter d from the system ; the diode.

es of the cascode high frequency e neutralization. own in Fig. 38, cted so that the reduces internal

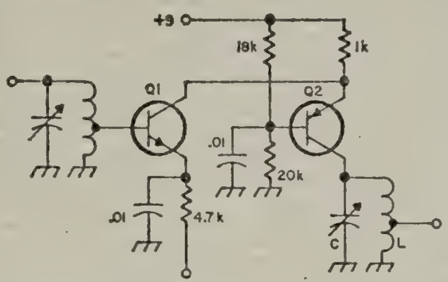


Fig. 38. This cascode amplifier is extremely useful because it provides high gain without the need for neutralization. Q1 is a 2N918, 2N3464, 2N3478, MPS918, 40235 or HEP-56; Q2 is a 2N1742, 2N2398, 2N2894, 2N3399, TIM10 or HEP-2.

feedback; therefore the input and output impedance of the circuit is essentially independent of the source and load.

In single transistor amplifiers the collector to base capacitance causes internal negative feedback that reduces amplifier gain at high frequencies. This feedback also causes the input and output impedances of the transistor to be dependent upon the value of the source and load impedances. Normally this negative feedback is neutralized out with a small amount of positive feedback external to the transistor. Unfortunately however, this process is long and tedious and often requires many adjustments.

The isolation between several similar stages is particularly important where more than one stage is used, because as a multi-

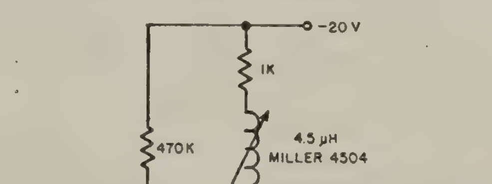


Fig. 39. This 30 MHz *if* stage uses two FET's connected in the cascode arrangement to provide 20 dB gain without neutralization; the bandwidth is 4 MHz. Both FET's in this circuit are 2N3819 MPF105 or TIS34. With a negative supply voltage, the 2N4360 or TIM12 would be suitable.

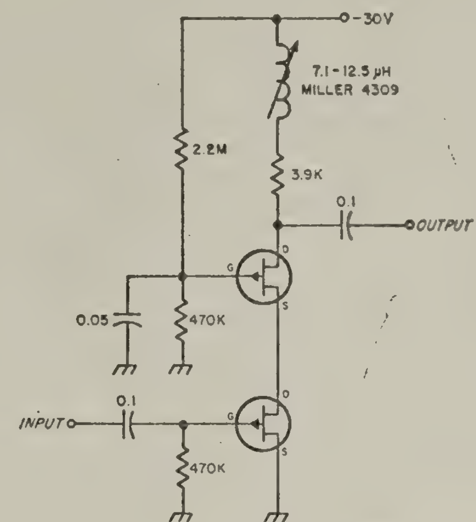


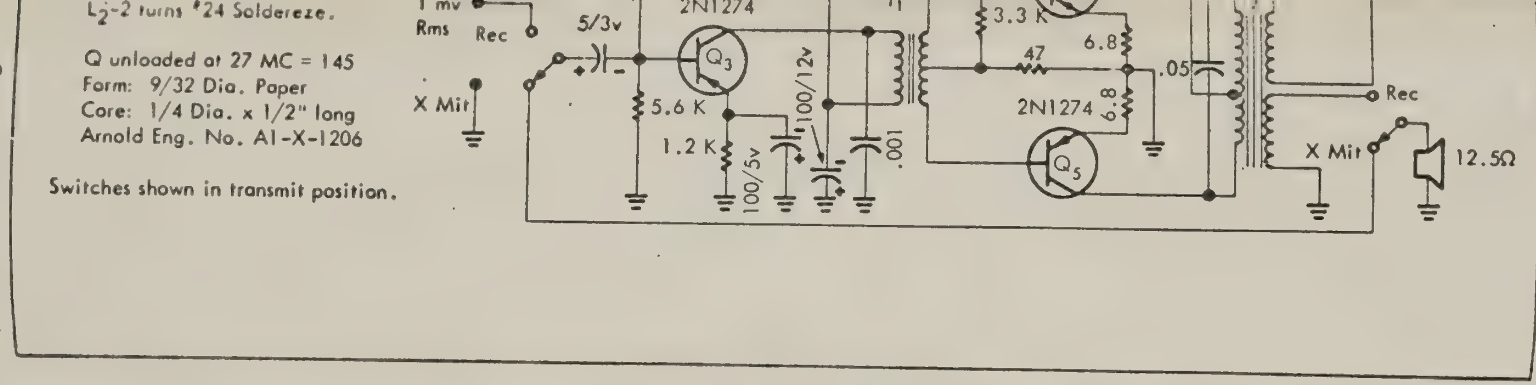
Fig. 40. This cascode video amplifier provides more than 6 MHz bandwidth with a voltage gain of 10. Both FET's are 2N3819, MPF105, TIS34 or HEP-801. A negative supply would permit the use of a 2N4360 or TIM12.

stage amplifier is being aligned, the tuning of one stage effects all the other stages. With the cascode rf amplifier, this shortcoming is overcome. Moreover, the gain of the cascode circuit is greater than the gain of a neutralized common-emitter stage with the same stability.

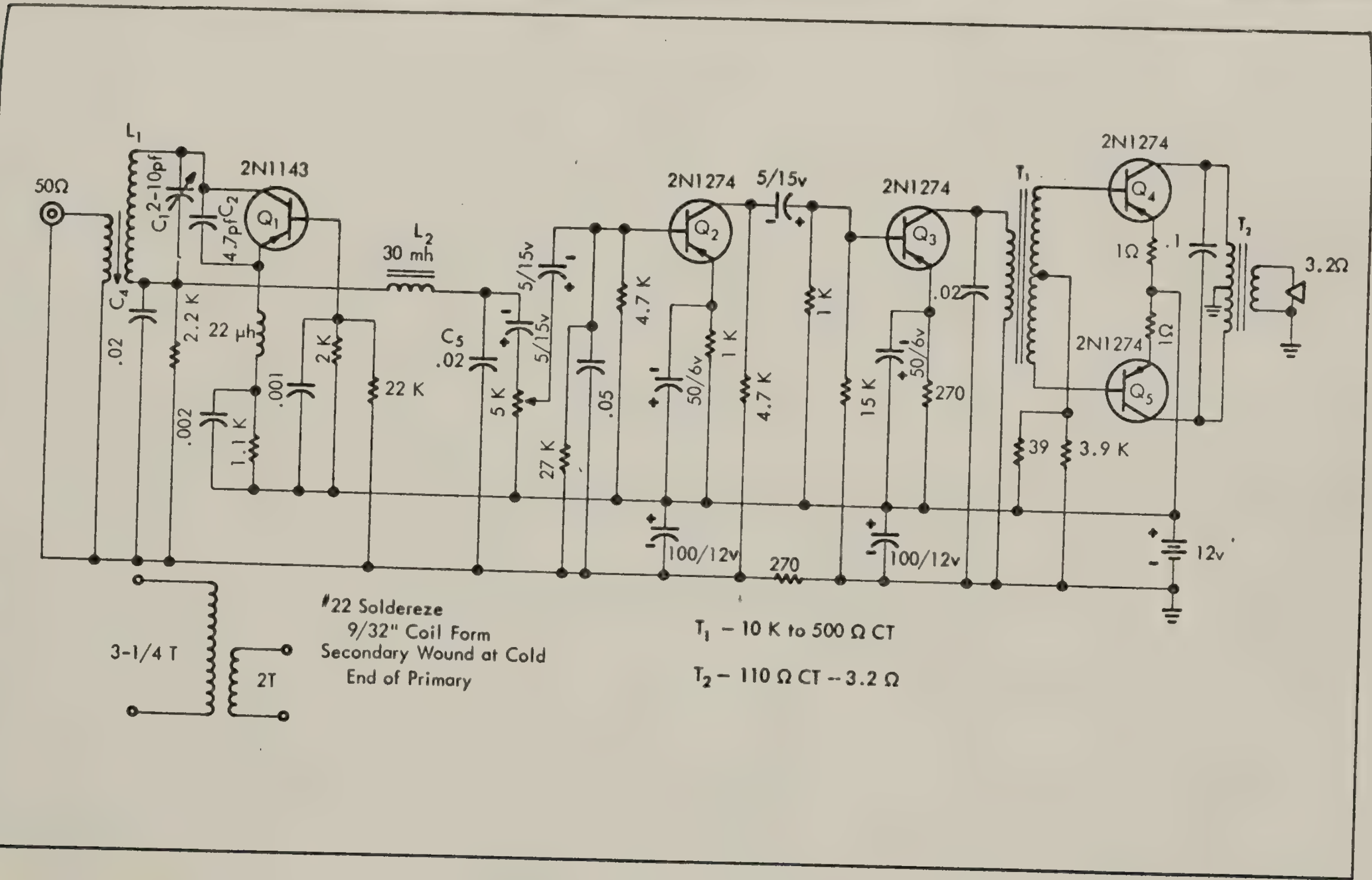
Since the emitter of Q1 is tied directly to the negative supply, the base can be connected directly to the output of the previous stage which is at ground potential. This eliminates a coupling capacitor and speeds up circuit recovery time after an overload. In addition, the gain of the stage may be controlled by varying the amount of current through Q1 (by adjusting the value of the negative supply).

The high gain cascode circuit shown in Fig. 38 uses two very inexpensive transistors to obtain 15 dB gain at 100 MHz with no neutralization; at lower frequencies the gain will be somewhat higher. No values are shown for the tuned circuits because they will be different for each application. However, dc biasing is usually the toughest part of any amateur transistor circuit design, and that is already done; all you have to do is put some tuned circuits in. The tap point on the input inductor should be chosen for best noise figure. The tap on the output inductor is chosen for maximum power gain.

Another somewhat different cascode circuit is illustrated in Fig. 39. This cascode 30 MHz *if* amplifier uses FET's and is



Courtesy Texas Instruments Inc



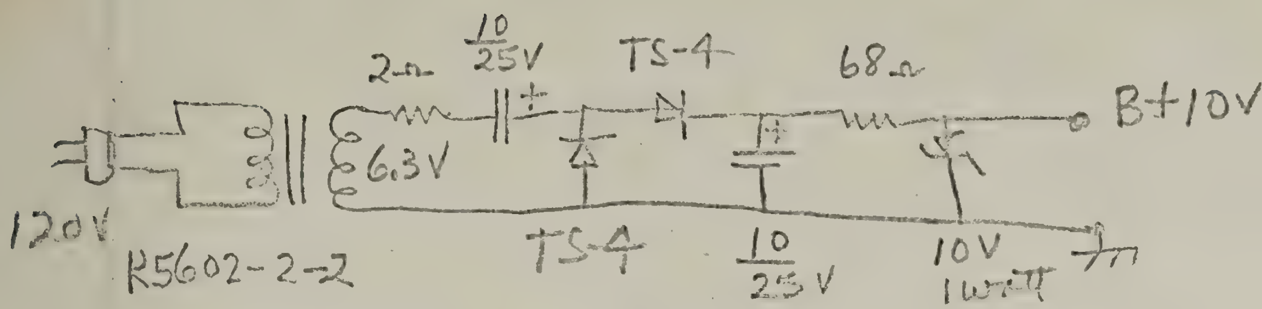
Courtesy Texas Instruments Inc.

Superregenerative 130-mc receiver.

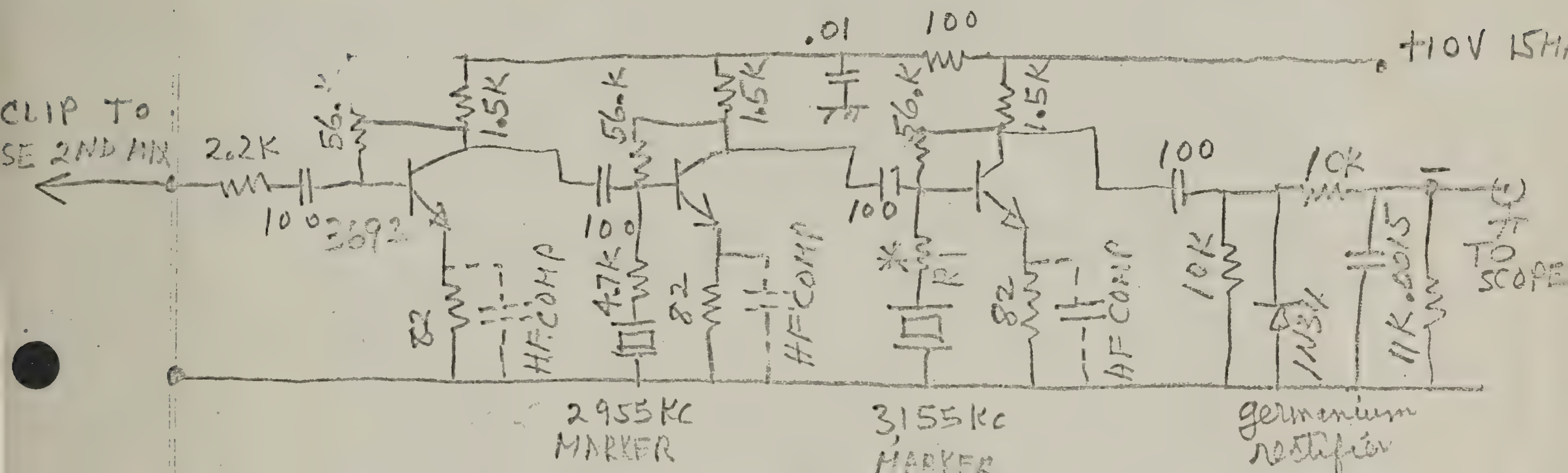
4uV sensitivity
BU 41-C

HQ 215 Video amplifier and Video detector 1-30-68

E. McDADE

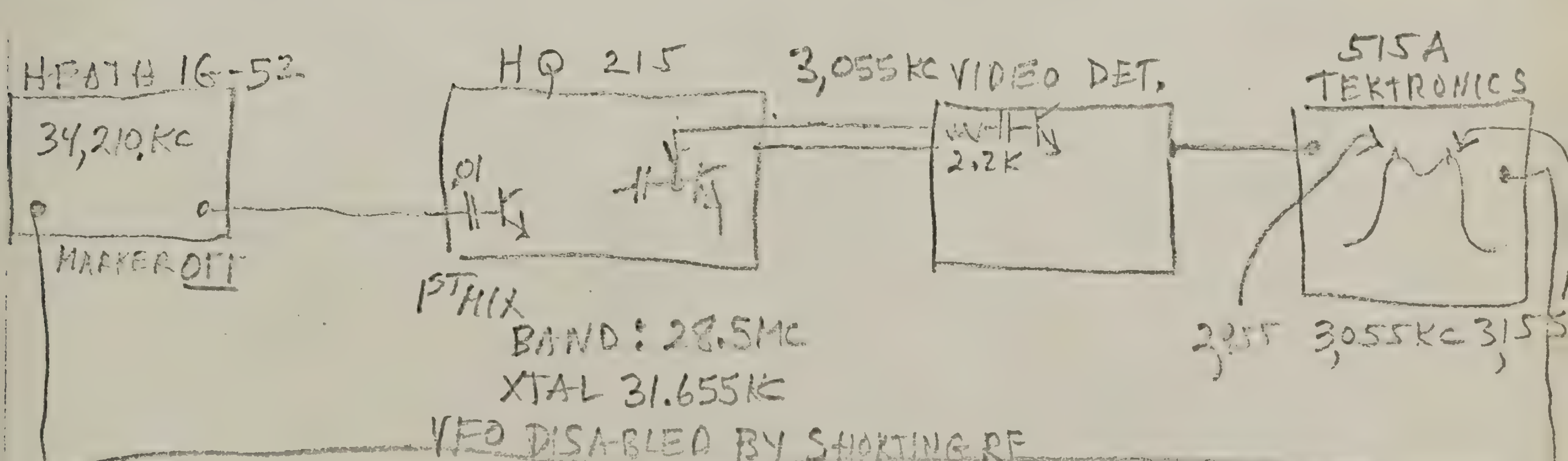


Voltage doubler,



VIDEO AMPLIFIER

* R1 was 470Ω, not 4.7K because 3155 xtal marker is weak



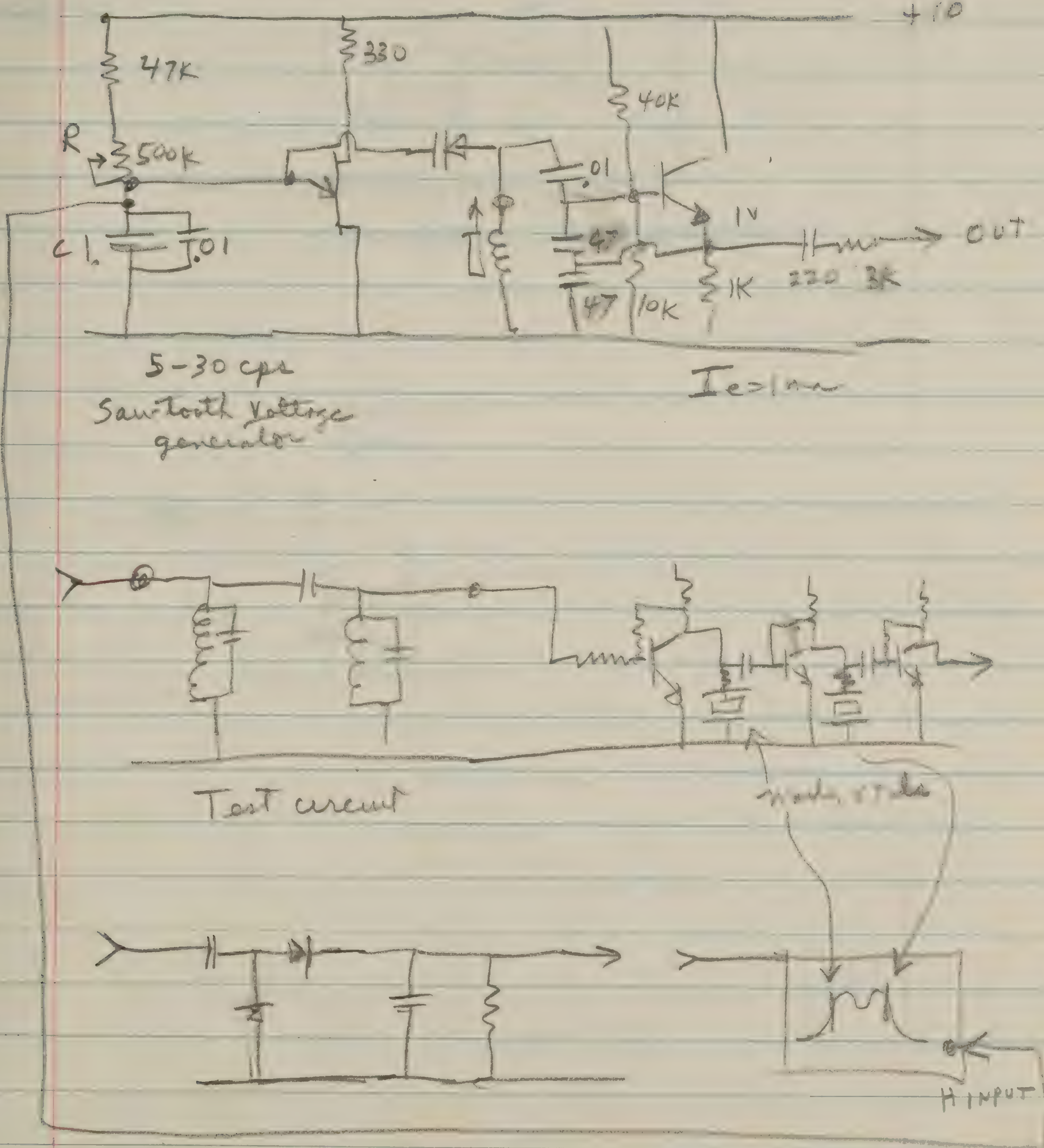
BAND: 28.5MC

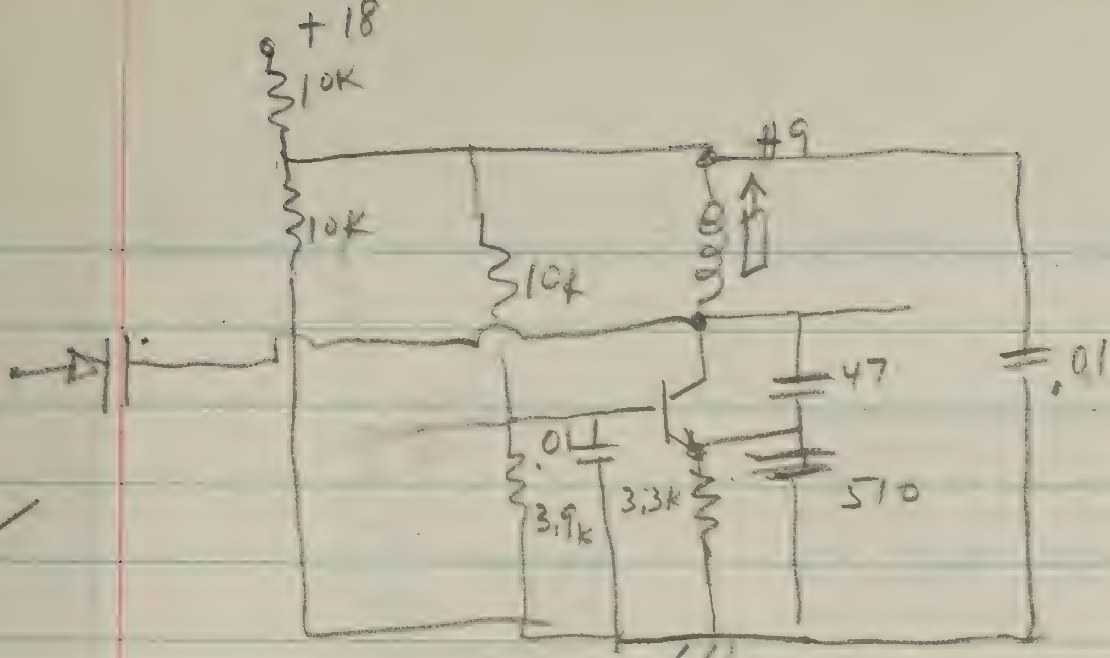
XTAL 31.655KC

VFO DISABLED BY SHORTING RF

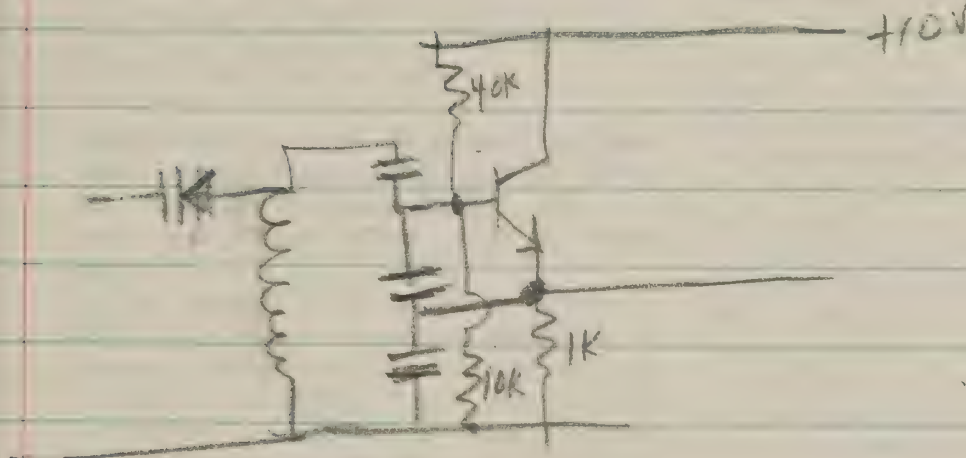
Sweep Frequency Generator

Ref 73 magazine p 28-29A





grounded Base, Colpitts oscillator



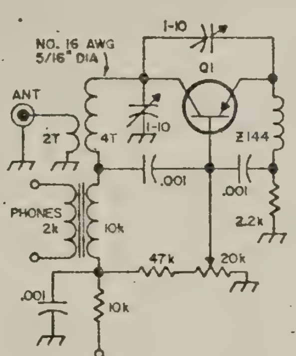


Fig. 41. This simple one transistor superregenerative receiver for two meters may be used for copying many local signals. With the components shown, this receiver will tune from about 90 to 150 MHz. It may be used on other frequencies by changing the inductor and capacitor as described in the text. Q1 is a 2N1742, 2N2398, 2N3399, TIM10, GE-9 or HEP-2.

characterized by a 20 dB power gain, bandwidth greater than 4 MHz, and all without the necessity for neutralization. For even more gain, these stages may be simply cascaded.

The cascode video amplifier shown in **Fig. 40** is almost an exact replica of the vacuum tube circuit that was originally developed in the 1940's. Usually a single triode is avoided in wideband or video amplifiers because the input capacitance and its multiplication by device gain (called Miller effect) seriously loads the input. With the cascode configuration, the input capacitance is no longer multiplied by the gain of the device, but is limited to the input capacitance of the FET. With the shunt peaking inductance shown (Miller 4309), this circuit is characterized by a voltage gain of 10 with a 3 dB bandwidth of over 6 MHz. The value of the inductance shown should work with most loads, but in some cases the bandwidth may be increased by changing its value slightly.

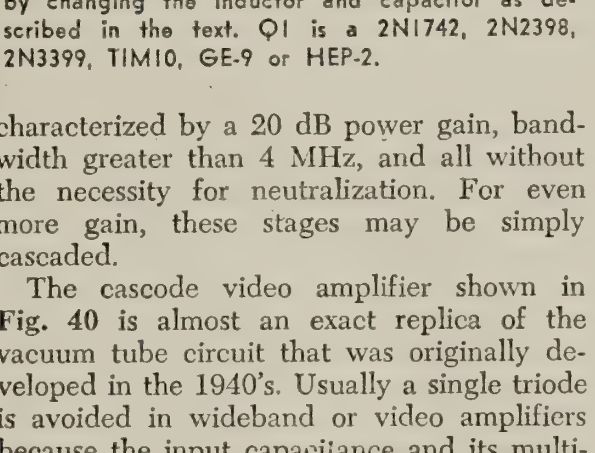


Fig. 42. Low noise two meter preamplifier uses a beer can cavity to provide excellent discrimination against nearby kilowatts. Q1 is a 2N3478, 2N3563, 2N3564, 40235 or SK3019.

Two meter superregenerative receiver

The simple little receiver shown in **Fig. 41** was designed primarily for the two meter band, but with appropriate changes in the input rf coils, it will work equally well on any frequency between 28 and 160 MHz. Basically, Q1 is a common base oscillating detector stage which quenches at about 25 kHz. The audio output from the detector is coupled across the transformer to the simple two stage audio amplifier. For more audio output, a 500 to 3.2 ohm transformer may be connected between Q3 and the speaker. The 20 kilohm pot should be adjusted for a total current drain of about 30 mA.

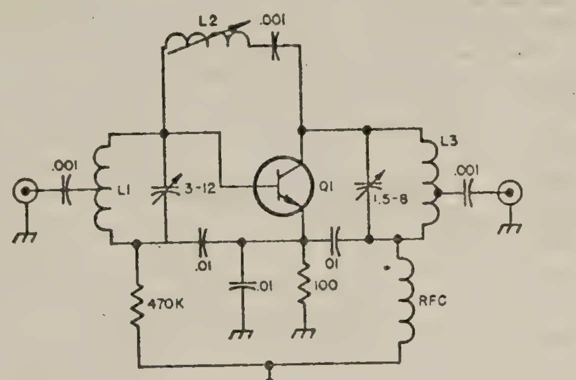
With the circuit constants shown in the diagram, this receiver will tune from about 90 to 150 MHz. To cover six meters, the tuning capacitor should be changed to a 15 pF unit and the number of turns on L2 increased to 6½. Since this is a regenerative receiver, it must be completely enclosed in a metal box to avoid the undesirable effects of hand capacity.

Two meter preamp

The two meter preamplifier of Fig. 42 uses inexpensive transistors in the common base configuration, yet provides noise figures that are nearly optimum for this band. The coaxial input cavity is easy to build and will provide more than adequate selectivity against your high powered neighbors on two. Although this input cavity may be built from scratch, the easiest approach is to use a spent beer can. Be careful when choosing the can though, some brewers are using aluminum cans which are pretty difficult to solder. The whole amplifier may be mounted on a small piece of epoxy board and then the entire assembly attached to the outside of the cavity.

220 MHz rf amplifier

The 220 MHz rf amplifier shown in Fig. 43 exhibits about 17 dB gain, although the exact amount will depend upon the type of transistor used. Optimum operation for low-



L1 3-1/2 TURNS NO. 16, 1/4" DIAM, 1/2" LONG. TAPPED AT CENTER.
L2 8-1/4 TURNS NO. 24 ON 1/4" SLUG-TUNED FORM.
L3 8 TURNS NO. 16, 1/8" DIAM, 7/8" LONG. TAPPED ONE TURN FROM COLD END.
Q1 2N3478, 2N3564, 40235
RFC 0.84 μH (OHMITE Z-220)

Fig. 43. Low noise 220 MHz preamplifier. This circuit will provide extremely high gain with low noise on the 1¼ meter band. Neutralization is controlled by inductor L2.

est noise figure also depends upon the transistor type, but usually will be in the vicinity of 6 volts collector voltage and 1 mA collector current. Unfortunately, this is usually not the same operating condition for maximum gain; maximum gain will occur at somewhat larger values of collector voltage and current. When this amplifier is set up, the 470k base bias resistor should be adjusted for the condition that you are looking for, whether it be minimum noise figure, maximum gain or a compromise between the two. Remember that for most practical receivers the noise figure will be dictated by the noise figure of the first rf stage; more gain can be added by another stage of amplification.

Circuits for Transmitters

Crystal oscillators

A compact untuned crystal oscillator is a very useful unit to have around the shack. The oscillator illustrated in Fig. 44 does not have any tuned circuits, so almost any crystal from 300 kHz up to 10 MHz will oscillate satisfactorily. It can be used for driving transmitters, as a signal source or for just testing crystals. In this circuit the first transistor is operating as an untuned crystal oscillator with the second transistor

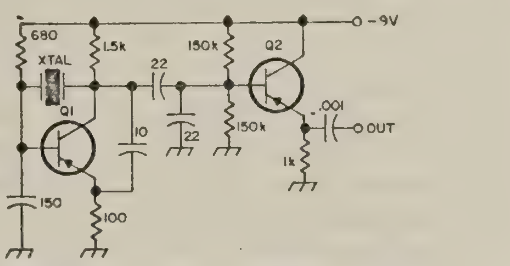


Fig. 44. This untuned crystal oscillator will oscillate with any crystal from 300 kHz to 10 MHz. Frequency stability is very good because the emitter follower buffer amplifier effectively isolates the oscillator from the load. Q1 and Q2 are 2N993, 2N1749, 2N2084, 2N2362, TIM10, GE-9, SK3006 or HEP-2.

connected as an emitter follower. With this arrangement, Q2 acts as a buffer stage and quite effectively isolates the oscillator from the load.

Another untuned crystal oscillator stage is shown in Fig. 45. This circuit will oscillate with any crystal between 3 and 20 MHz with no tuning whatsoever. If overtone crystals are plugged into the circuit, they will oscillate on their fundamental frequency. For overtone crystals up to about 60 MHz, the fundamental will be *approximately* ⅓ the marked frequency; above 60 MHz the fundamental is normally about ⅕ the marked frequency. For best stability with each of these untuned crystal oscillators, all the capacitors should be high grade silver mica types.

The crystal oscillator shown in Fig. 46 has proven to be extremely stable and easy to adjust. Basically it is a standard Colpitts circuit with the frequency determined by the crystal. By using the appropriate inductors and capacitors, this circuit will oscillate

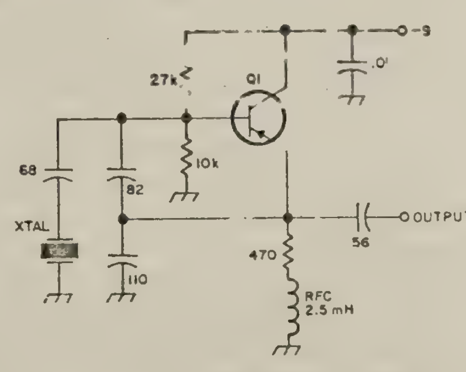
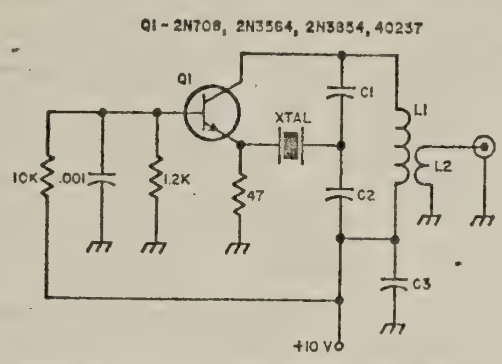


Fig. 45. This crystal oscillator will oscillate with any crystal between 3 and 20 MHz with no tuning whatsoever; overtone crystals will oscillate on their fundamental in this circuit. Q1 is a 2N1177, 2N1180, 2N1742, GE-9, SK3006 or HEP-2.



FREQUENCY	C1	C2	C3	L1	MILLER NO.	L2
10 - 13.5 MHz	30pf	300 pf	.01	5.0 - 9.0 μH	4505	4-1/2T
13.5 - 18 MHz	30pf	300 pf	.01	2.8 - 5.0 μH	4504	3T
18 - 24 MHz	30pf	300 pf	.01	1.6 - 2.8 μH	4503	2-1/2T
23.5 - 32 MHz	10pf	100 pf	.01	2.8 - 5.0 μH	4504	3T
32 - 42 MHz	10pf	100 pf	.001	1.6 - 2.8 μH	4503	2-1/2T
42 - 53 MHz	10pf	100 pf	.001	1.0 - 1.6 μH	4502	2T
58 - 84 MHz	10pf	100 pf	.001	0.4 - 0.8 μH	4501	1-1/4T

Fig. 46. This Colpitts type crystal oscillator may be used with either fundamental or overtone crystals from 10 MHz to 84 MHz with the tuned circuit components listed. It oscillates quite readily when adjusted and provides a stable output.

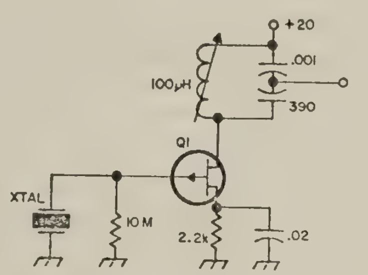


Fig. 47. This is the old familiar vacuum tube Pierce oscillator circuit with a field effect transistor in place of the thermionic triode. Circuit constants shown here are for the 1 MHz region, but the tuned circuit may be adjusted to any frequency desired. Q1 is a 2N4360 or TIM12.

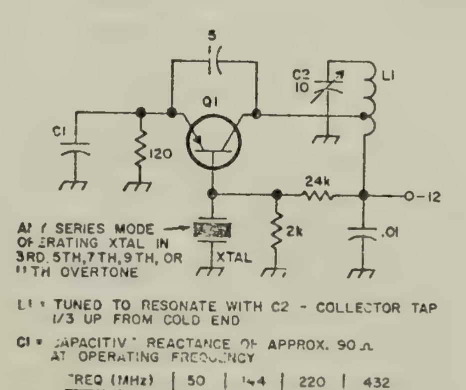


Fig. 48. This crystal oscillator was designed specifically for overtone crystals and will oscillate up to the 11th overtone in the VHF range. Suitable values for C1 are shown for the VHF bands; for other frequencies, C1 should exhibit approximately 90 ohms capacitive reactance for best results. Q1 is a TIM10, T1400 or HEP .

with either fundamental or overtone crystals. Although circuit values are only provided here up through 84 MHz, this circuit will operate well above 100 MHz with smaller values of capacitance; the only requirement is that they retain a 10 to 1 ratio in capacitance. For operation with a negative supply voltage, ground the 10 volt line shown in the schematic, lift the 1.2k and 47 ohm resistors from ground and tie them to the negative supply. PNP germanium transistors may also be used by reversing the supply voltage and changing the 10K base bias resistor to 33k.

The untuned crystal oscillator in **Fig. 47** uses an FET in the familiar Pierce vacuum tube circuit. In this oscillator the drain to source capacitance and gate to source capacitance make up the feedback path with the amount of oscillator excitation determined by their ratio. This circuit cannot be used with conventional junction transistors because their low input impedance severely loads down the crystal.

The crystal oscillator shown in **Fig. 48** is designed specifically for overtone crystals and will work up through the eleventh overtone. The circuit is completely noncritical except for the value of C1 which should exhibit approximately 90 ohms capacitive reactance at the operating frequency. The tuned circuit is tuned to the frequency of interest. The 5 pF capacitor from collector to emitter should be adjusted for maximum rf output; above about 200 MHz it may not be required. The constants shown in the schematic should cause oscillation with any overtone crystal in the VHF range, but in some cases a sluggish crystal may require adjustment of the 24k base bias resistor to take off every time power is applied.

Variable crystal oscillator

The variable crystal oscillator shown in **Fig. 49** is a very useful circuit to the ham who wants a highly stable signal on two meters or 432. Although it will only tune about 50 kHz on two and 150 kHz on 432, it is adequate for many types of operation. On 432 for example, most operation is within a few kHz of 432.00 MHz. The circuit's operation is quite straightforward; the dual 365 pF capacitor pads down the resonant circuit and *pulls* the crystal *down* in frequency. Just how much it is pulled down is determined by the inductor L1. For an

overtone crystals.
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this circuit will
Hz with smaller
only requirement
1 ratio in capaci-
a negative supply
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2k and 47 ohm
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scillator shown in
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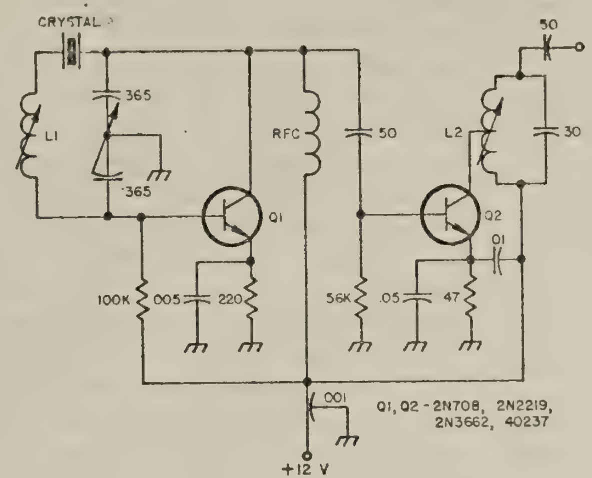


Fig. 49. This variable crystal oscillator (VXO) may be used to vary the frequency of an 8 MHz crystal 4 or 5 kHz when the 365 pF dual variable is tuned through its range. When multiplied to two meters or 432, this provides a very stable variable frequency. For 8 MHz crysals, L1 is a 20-25 uH slug tuned coil; L2 is chosen to resonate at 8 MHz with the 30 pF capacitor.

Table 1

Crystal	L_1		L_2*
3.5 MHz	35-60 μH	Miller 4509	80 turns #36, tapped at 27 turns
5.0 MHz	24-35 μH	Miller 4508	62 turns #36, tapped at 21 turns
8.0 MHz	16-24 μH	Miller 4507	40 turns #36, tapped at 13 turns
9.0 MHz	16-24 μH	Miller 4507	36 turns #36, tapped at 12 turns

*Wound on ¼" slug tuned form.

8 MHz crystal, this inductor should have a center value of about 22 μH; it should exhibit relatively high Q at 8 MHz and be self resonant well above the crystal frequency. As this inductor is increased beyond a certain amount, the crystal will lose control and the circuit becomes a rather inferior VFO. For best results L1 should be adjusted so that the crystal is pulled 4 or 5 kHz when the variable capacitor is tuned through its full range.

The buffer amplifier is coupled to the oscillator through a 50 pF capacitor. For maximum frequency stability, this capacitor should be the minimum value that will provide adequate drive for your transmitter. With the 50 pF capacitor shown, approximately 10 volts of 8 MHz drive should be available with the buffer tank tuned to resonance. Inductor L2 is chosen to resonate at 8 MHz with the 30 pF capacitor; the tap is ⅓ up from the ground end.

Two frequency crystal oscillator

In the two frequency crystal oscillator illustrated in Fig. 50, the bilateral characteristics of the transistor effectively provide two separate common emitter stages. Either of the two frequencies may be selected by simply applying a positive or negative voltage to the circuit.

When a positive voltage is applied, current flows through D1 to the emitter of the transistor. The tuned circuit consisting of L2, C2 and the crystal Y2 determine the oscillation frequency available at the output. The other tuned circuit consisting of L1 and C1 is shorted out by D1. In addition, since crystal Y1 is connected between the base and emitter, there is no gain to promote oscillation at its frequency.

If a negative voltage is applied to the supply terminal, the transistor "inverts" itself with the collector becoming the emitter and the emitter the collector. In this case L1, C1 and Y1 determine the frequency of oscillation. Diode D2 shorts out the other tuned circuit and the crystal Y2 is connected between the base and emitter of the inverted transistor; therefore, there is no output at Y2's frequency.

Transistors may not normally be used in the inverted mode because rather large

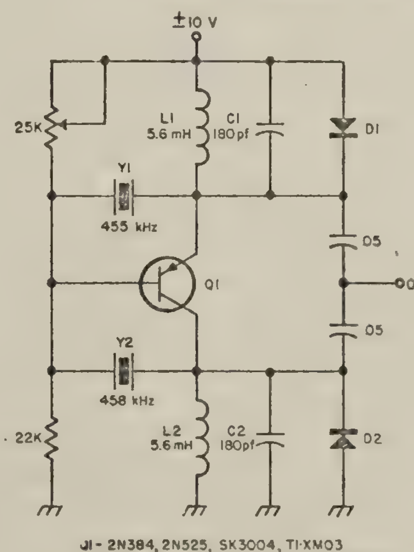


Fig. 50. This two frequency crystal oscillator changes frequency by simply reversing the supply voltage. When the supply voltage is changed, the transistor inverts itself; usually transistors may not be used in the inverted mode, but in an oscillator a gain of only 1 or 2 is needed and this circuit provides a novel and simple way of obtaining two frequencies from a single stage with a minimum of switching.

amounts of gain are desired. However, as an oscillator, the gain need only be sufficient to produce oscillation; this usually requires a forward current gain of only one or two. For this reason almost any germanium transistor may be used in this application. Silicon NPN transistors will also work, but operation will be just opposite to that described above.

Diodes D1 and D2 limit the output voltage to about 0.7 volts, so for some applications, further amplification may be necessary. The tuned circuit values shown in the schematic are for a resonant frequency of 455 kHz, where this circuit provides an excellent method for upper and lower sideband selection. It may be used on other frequencies by simply changing the values of inductance and capacitance in the tuned circuits.

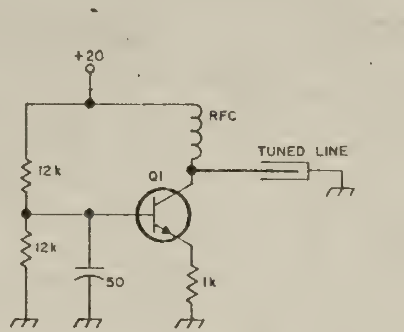


Fig. 51. This simple UHF oscillator will provide about 2 mW up to 1000 MHz; some selected transistors will provide usable power up to 1500 MHz or so. Q1 is a 2N918, 2N3478, 2N3564 or HEP-56.

UHF oscillator

The simple UHF oscillator circuit shown in Fig. 51 will deliver up to about 2 mW of power at 1000 MHz. Although this amount of power is insufficient for some applications, 2 mW is more than enough for many mixer and converter circuits. Many transistor types, when selected, will oscillate up to 1500 MHz in this simple circuit.

Ten meter transmitter

The three watt ten meter transmitter shown in Fig. 52 gets over the high rf power/high price hurdle by using three inexpensive transistors in parallel in the final stage. The three paralleled transistors used will produce three watts output with a 15 to 18 volt supply and about 2.25 watts with a 12 volt supply. The rf drive is provided by a 28 MHz crystal oscillator and driver amplifier. For maximum efficiency, modulation is applied to both the final amplifier and driver through the modulation transformer; about 1.5 watts of audio power is required for 100% modulation. Since the transistors used in this transmitter have an f_T of 500 MHz, a similar transmitter could be built for six meters; the only change would be in the resonant circuits.

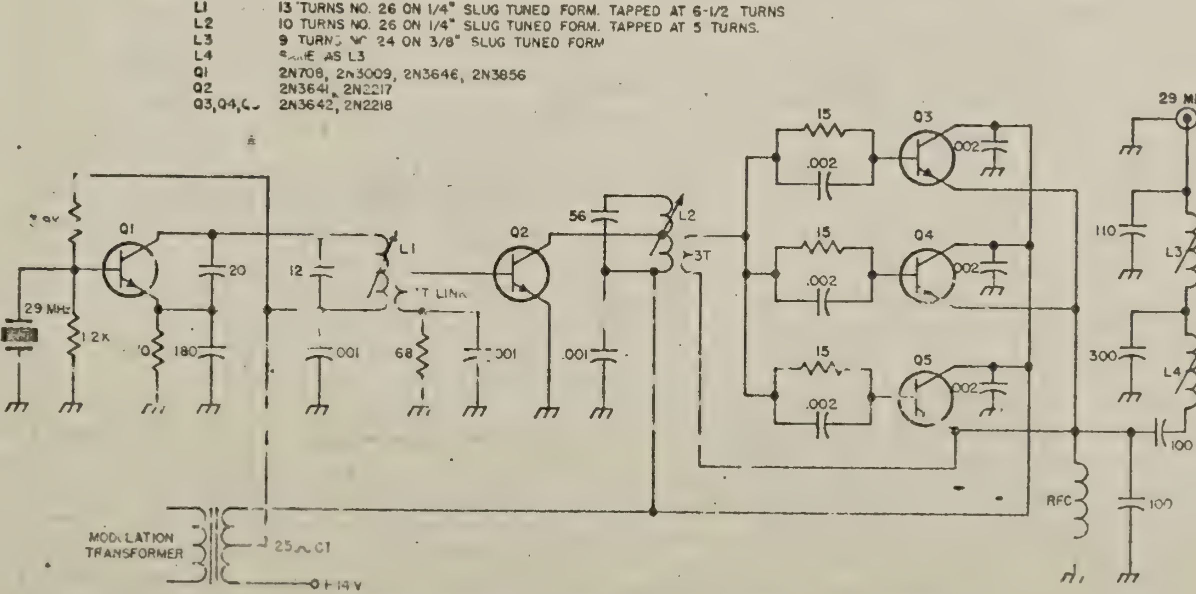
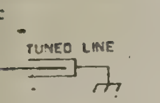


Fig. 52. This three watt ten meter transmitter maintains high efficiency and low cost by paralleling three inexpensive silicon transistors in the final stage.



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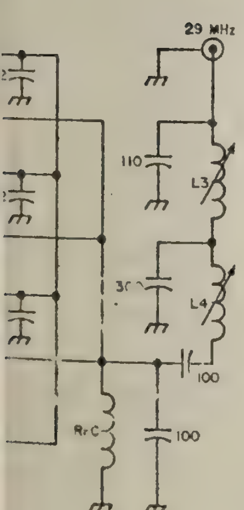

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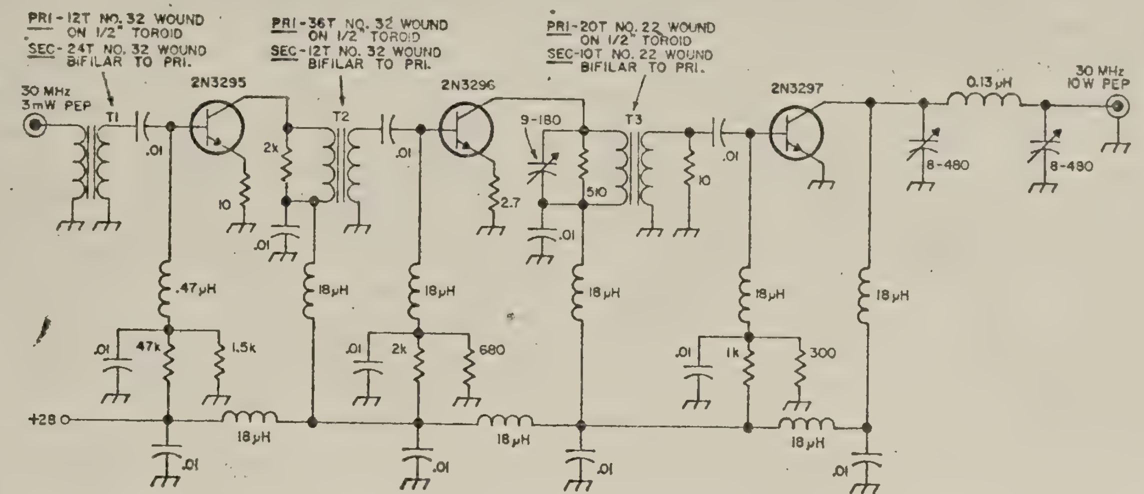


Fig. 53. This ten meter linear amplifier for SSB service uses transistors which were designed specifically for single sideband linear operation. Many junction transistors cannot be used satisfactorily for this application, because linear amplification at low power levels is a serious problem.

Ten meter linear amplifier

Up until the present time transistors haven't been used too much in SSB transmitters because linear amplification at even low signal levels has been a serious problem. However, the transistors in the ten meter SSB power amplifier illustrated in Fig. 53 were designed specifically for linear amplifier service and perform quite well. The measured distortion of these devices is less than three percent without feedback, which is somewhat better than tubes under the same conditions.

Actually the circuit of this amplifier is quite straight forward. The only critical parts are the coupling transformers between succeeding stages. These are wound on small ½" toroids which are suitable for use at 30 MHz (Ami-tron T-50-2). Coupling between stages must be very tight and the transformers should be bifilar wound. Both the input and output of this unit are designed for 50 ohm coaxial line, so it fits in nicely with other equipment being used on ten meters.

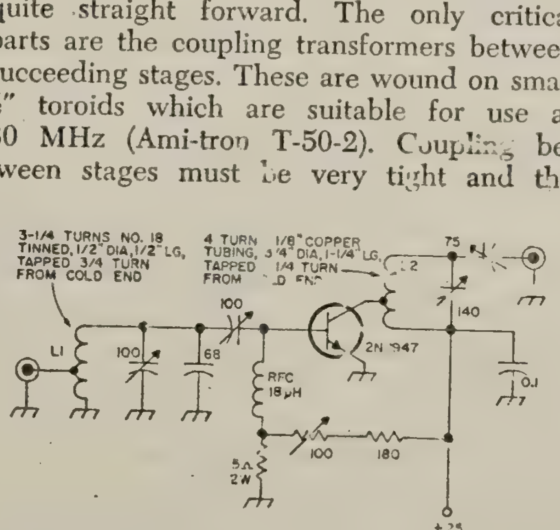


Fig. 54. This ten meter single sideband linear power amplifier will provide up to 8 watts PEP. The power gain of the 2N2947 is 13 dB at this frequency, and the odd order distortion products are at least 30 dB below the desired output.

The ten meter single sideband linear power amplifier shown in Fig. 54 is capable of delivering an output power of 8 watts PEP. The power gain at this frequency is 13 dB, and all odd-order distortion products are at least 30 dB below the desired output.

The main difference between this amplifier and one designed for class C operation in CW, AM or FM transmitters lies primarily in the dc bias circuit. For class C operation, the only dc bias normally applied is the collector supply voltage. The 18 μH rf choke and resistive divider in the base circuit would be omitted. The transistor is biased *on* by the driving signal on the base. This results in one of the big advantages of the transistor transmitter—if the driving signal is suddenly removed, the power amplifier merely shuts off and sustains no damage.

To obtain linear operation, a small amount of forward bias is applied to the transistor. This is a function of the resistive divider and the isolating choke in the base circuit. The bias is adjusted so that a small collector current flows without any input driving signal; when a driving signal is applied, the transistor is biased on to full operating collector current. In this circuit the 2N2947 draws 20 mA with no drive and 350 mA with full drive.

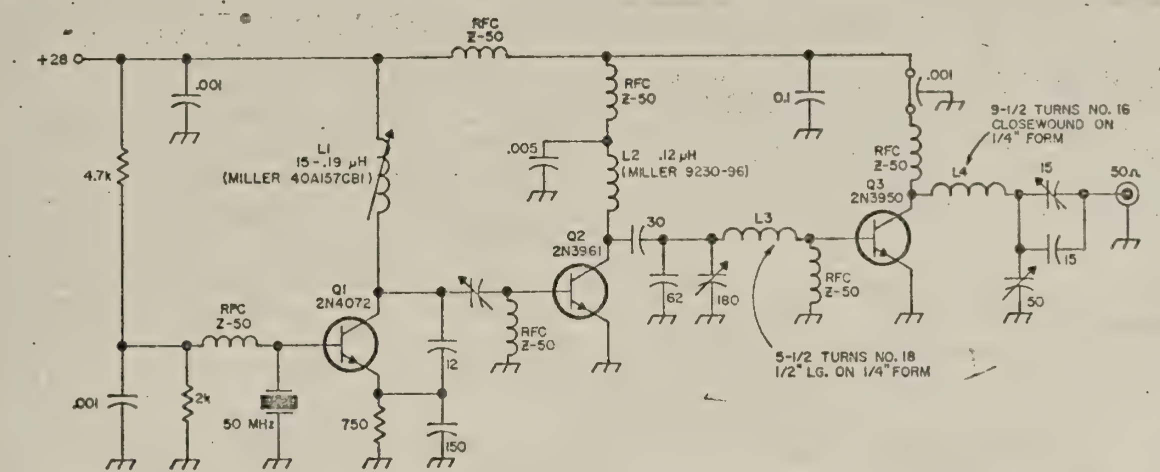


Fig. 55. This 6 meter transmitter provides up to 50 watts of power with very good efficiency and very low harmonics. The 2N3950 in the final provides a minimum power gain of 8 dB on six and is rated at 50 watts continuous service.

Six meter transmitter

The six meter transmitter of Fig. 55 will provide 50 watts of power into the antenna with very good efficiency and very low harmonics. The second harmonic is suppressed on the order of 28 dB while the third harmonic is more than 34 dB down. The efficiency of the final stage is 69% and the overall efficiency of the entire transmitter is 62%. The bulk of the total current drain of 2.9 A is required by the final amplifier —2.6 A. By choosing each of the circuit components very carefully, a transmitter evolved which uses only three transistors where several more stages are normally required. The mainstay of this transmitter however is the 2N3950 transistor in the power amplifier; this transistor can provide 50 watts of continuous power at 50 MHz with a minimum power gain of 8 dB.

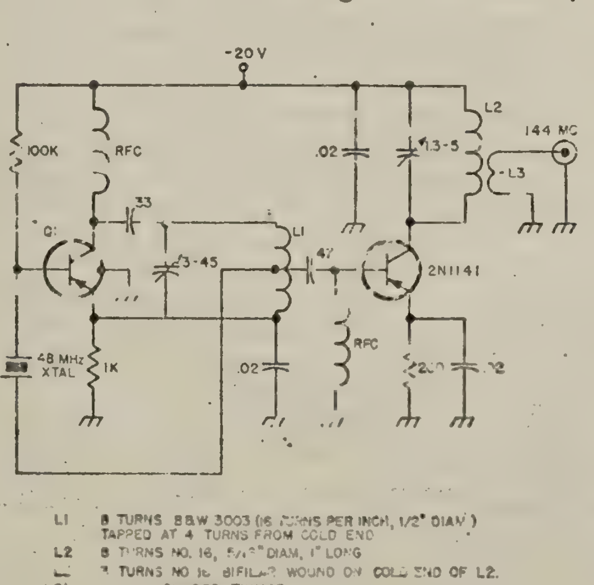


Fig. 56. This simple two meter transmitter may be used as a driver for a larger 144 MHz transmitter or a signal source for testing receivers, converters and antennas.

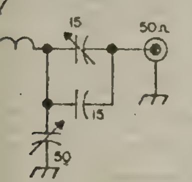
Two meter driver

The simple two meter driver shown in Fig. 56 is just about the minimum that is suitable for driving a small 144 MHz transmitter. The first stage of this VHF driver consists of a crystal controlled oscillator operating at 48 MHz; the 48 MHz output from the oscillator is capacitively coupled to the 2N1141 tripler stage tuned to 144 MHz. The two meter output of this circuit is quite low, but sufficient to drive a small power amplifier to a quarter watt or so. This circuit may also be used as a two meter signal source, or as a source for a VHF SSB mixer.

Circuits for Test Equipment

Signal tracer

The signal tracer is a universally used unit of test equipment which may be used for troubleshooting and isolating defective stages in all types of electronic equipment. With a suitable rf probe it may be used to check the operation of rf and *if* amplifiers; with an audio pickup probe it may be used as a straight through audio ampli-



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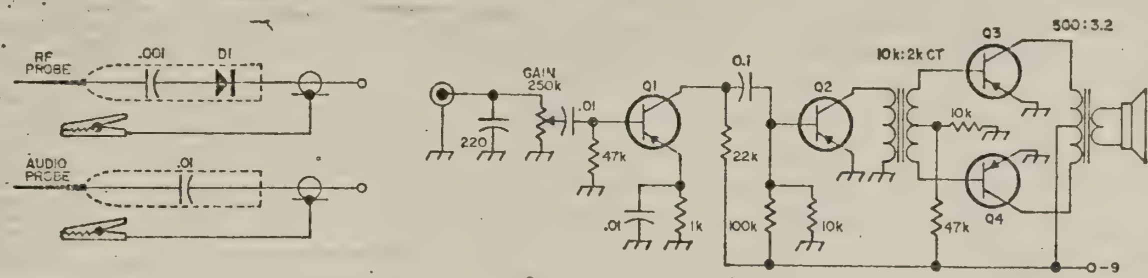


Fig. 57. This signal tracer provides more than adequate audio output with only 100 microvolts of modulated rf at the input. It may also be used for tracing audio circuits, but don't depend on its fidelity. All the transistors are germanium types such as the 2N404, 2N1450, 2N2953, SK3004, GE-2 or HEP-253; the diode in the probe is a 1N34A or 1N67A or similar.

fier to check microphones and preamps and to detect noise and hum in amplifiers.

The signal tracer shown in Fig. 57 uses a push pull output stage which produces more than adequate audio output to a miniature speaker with only 100 microvolts input. The output from the probe is applied directly to the 250 kilohm pot, which serves as a gain control and as a diode load when an rf probe is being used. The signal is coupled to the base of the first transistor, amplified, and fed to the driver stage and transformer coupled push pull output.

Only a very small amount of audio signal is necessary to operate the signal tracer as a straight through audio amplifier. However, don't use it to check fidelity because it is designed primarily for maximum sensitivity without regard to frequency response.

When using this signal tracer always start with the audio gain control turned all the way down because it is easy to overload the simple amplifier; the result is a highly distorted output signal. In some receivers the rf probe may load the mixer plate or *if* grid. If this happens, a tone modulated signal should be injected at the antenna terminals of the receiver to obtain a usable output from the signal tracer.

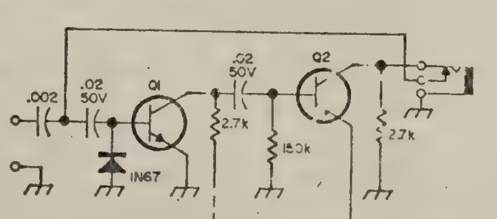
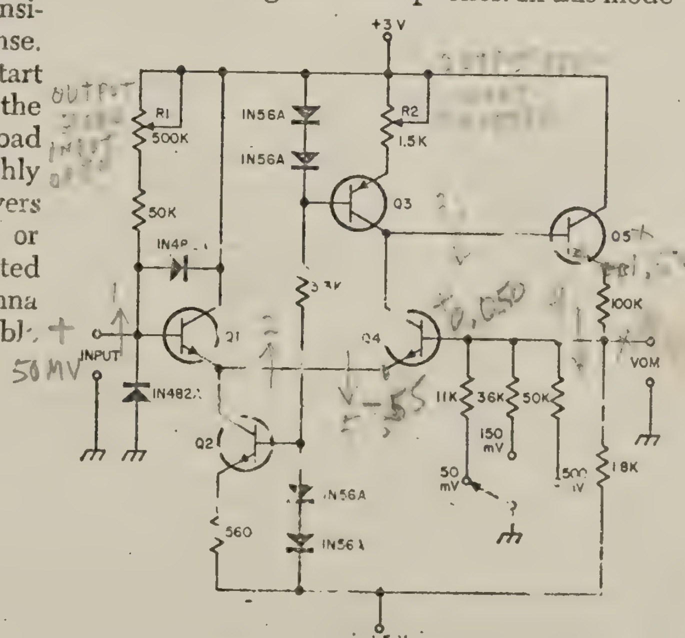


Fig. 58. This signal injector/tracer switches from the injection mode to a signal tracer by simply plugging in a pair of high impedance magnetic earphones. As a tracer is works from audio up to 432 MHz. Transistor Q1 is a 2N1170, 2N388A, 2N1605, SK3011 or GE-7; Q2 is a 2N138A, 2N404, 2N2953, SK3004 or HEP-253.

Signal injector/tracer

The circuit illustrated in Fig. 58 functions as both a signal injector and signal tracer. Furthermore, no switching is required; it's all accomplished automatically when the headphones are plugged in for signal tracing. In the inject mode the circuit is a clamped multivibrator with extremely narrow pulses and high harmonic content. In fact, with this circuit, sufficient output is available for signal tracing from audio (750 Hz) to well above 40 MHz. This frequency range is more than adequate for most requirements.

The unit is switched to the signal tracer mode by simply plugging in a pair of high impedance magnetic headphones. In this mode



Q1,Q2,Q4,Q5 - 2N1304, 2N2924, 2N3393, SK3011
Q3 - 2N404, SK3003, SK3004

Fig. 59 This VOM range extender increases the sensitivity of your volt-ohm-milliammeter to 50 millivolts full scale; full scale readings of 150 and 500 millivolts are also provided by the range switch.

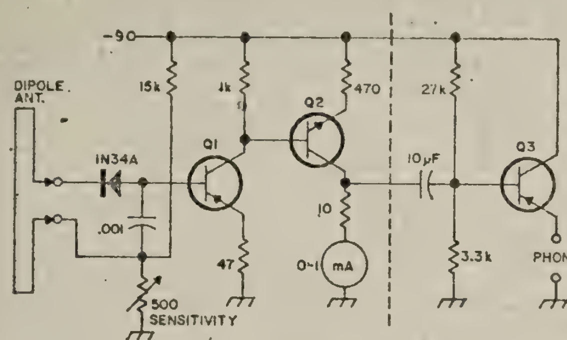


Fig. 60. This monitor/detector may be used for measuring field strength, monitoring modulation or finding hidden dipole transmitters by simply attaching a resonant dipole antenna at the input. Q1 and Q3 are 2N408, 2N2953, SK3004, GE-2 or HEP-253; Q2 is a 2N170, 2N388A, 2N1605, SK3011 or GE-7.

it will detect and amplify signals from 20 Hz to above 432 MHz. Since this circuit only requires about 100 microamperes, no on-off switch is provided. This very small current drain insures that the life of the battery will be nearly that of its shelf life. By using miniature components and a little care in layout, it is possible to mount this complete injector/tracer in an old penlight case or metal cigar tube.

VOM range extender

Most volt-ohm-milliammeters are not too suitable for use with transistor circuits because their lowest voltage scales are either 1.5 or 3 volts full scale and they are not sensitive enough to accurately measure the base to emitter voltage of a transistor which may be 120 millivolts or so. The low voltage dc preamplifier shown in **Fig. 59** is inexpensive, stable with temperature and supply voltage variations yet extends the range of any VOM so it can be used effectively in semiconductor circuit measurements.

In this circuit transistors Q1 and Q4 constitute an emitter coupled amplifier; Q5 is an emitter follower connected so the circuits entire output voltage is fed back to Q4. Transistors Q2 and Q3 are constant current sources in the negative and positive lines respectively. These constant current sources reduce the sensitivity of the amplifier to voltage supply variations and result in substantially lower drift. To control the gain of the amplifier for different voltage ranges a portion of the output voltage is fed back to the base of transistor Q4 through the voltage divider selected by the range switch. With the values shown in the schematic, this circuit provides gains of 3, 10 and 30, which extend the 1.5 volt scale of the VOM to 500, 150 and 50 millivolts full scale.

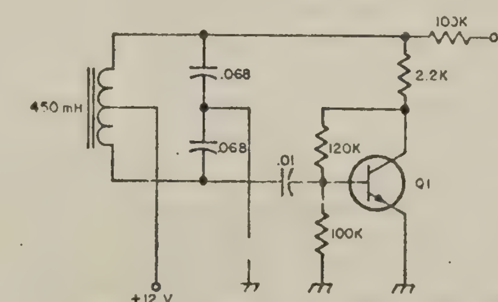
There are two zero controls which must be adjusted when using this unit; first the 500 kilohm pot (R1) in the base bias lead to transistor Q1 is adjusted to zero the output with no input and the base isolated from ground. The 1.5k zero adjust pot (R2) is then adjusted for an output zero with the input leads shorted together.

Monitor/detector

The simple VHF monitor/detector illustrated in Fig. 60 may be used for measuring field strength, monitoring modulation or even in hidden transmitter hunts. The frequency of use may be simply changed by changing the length of the dipole antenna. In some cases where the rf field is strong enough, a simple vertical pickup antenna will be sufficient for signal monitoring purposes. Furthermore, the use of this monitor/detector is not limited to the VHF bands; the addition of an appropriate antenna will permit its use at any frequency up to about 500 MHz. For lower frequencies where the size of the dipole antenna would be impractical, a simple vertical pickup antenna and tuned resonant circuit may be substituted at the input.

1 kHz oscillator

The simple 1 kHz oscillator shown in **Fig. 61** is very useful for many testing devices around the shop. This circuit is simply a Colpitts oscillator, but the circuit values and feedback have been chosen for maxi-



Q1 - 2N2924, 2N3393, 2N3566

Fig. 61. Simple 1000 Hz oscillator is very useful for testing and measurement around the shack. The Colpitts circuit is used with component values chosen for maximum stability and good waveform.

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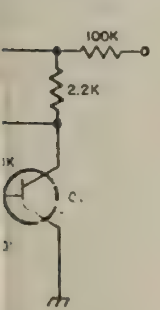

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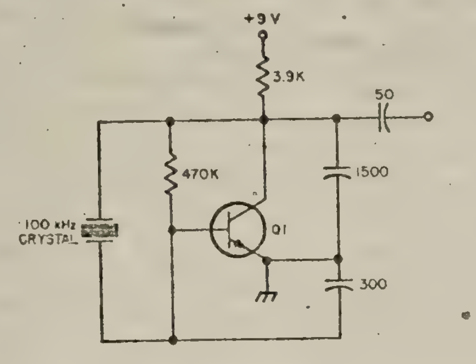


Q1 - 2N708, 2N2711, 2N2923, 2N3691

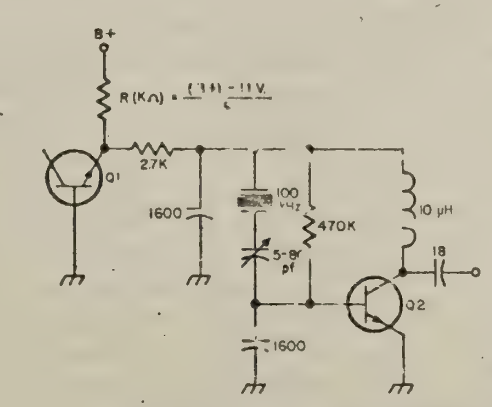
Fig. 62. The 100 kHz calibrator shown here is just about the simplest circuit that will provide usable results. For zeroing in with WWV a small padder capacitor may be added in series with the 100 kHz crystal.

mum stability along with good waveform. It may be used for testing speech and modulation equipment, or even as a driver for a code practice oscillator.

100 kHz crystal calibrators

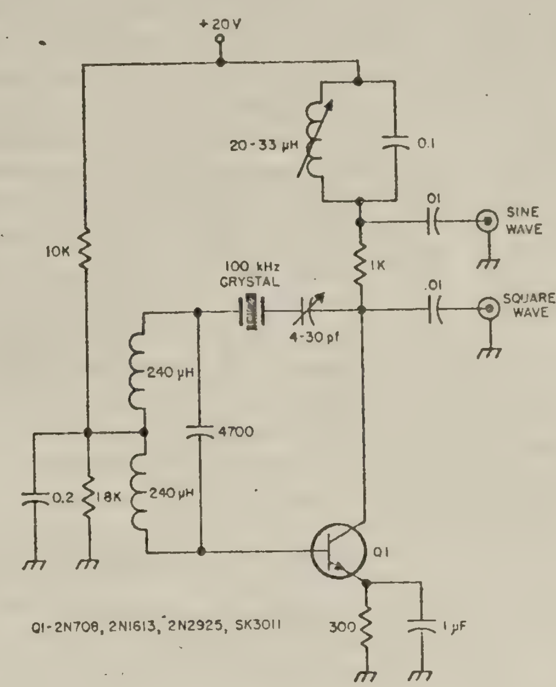
The 100 kHz crystal oscillator illustrated in Fig. 62 is just about as simple as you can build and still get a usable output. This circuit will provide usable harmonics up to about 30 MHz but it has no provision for zeroing in with WWV. This feature may be added by simply placing a small variable padder capacitor in series with the 100 kHz crystal.

The calibrator circuit shown in Fig. 63 is only slightly more complicated than its counterpart in Fig. 62, but provides usable harmonics up to 150 MHz and has a built in voltage regulator. In this case the base



Q1 ANY NPN PLANAR SILICON TRANSISTOR
Q2 2N2923, 2N3392, 2N3565, SE4002

Fig. 63. This 100 kHz crystal calibrator is only slightly more complicated than the one shown in Fig. 62, but has a built in voltage regulator (Q1) and provides usable harmonics up to 150 MHz.



Q1-2N708, 2N1613, 2N2925, SK3011

Fig. 64. This 100 kHz crystal calibrator uses a crystal in the parallel mode and provides either a sinusoidal or square wave output. The calibrator may be zeroed to WWV with the 4-30 pF trimmer.

to emitter junction of an NPN silicon planar transistor is connected as a zener diode. When connected in this manner, these transistors provide a regulated voltage of approximately 11 volts. The value of the series dropping resistor may be determined by using the formula shown in the schematic. The 5–80 pF capacitor is for zeroing in with WWV.

The 100 kHz crystal controlled oscillator shown in Fig. 64 uses a low cost silicon transistor, and provides both a square wave and sine wave output with excellent frequency stability. The oscillator circuit is basically the Hartley type with positive feedback from the collector to base through a phase reversal in the tapped tank circuit, consisting of two 240 µH chokes and a 4700 pF silver mica capacitor. The oscillator frequency is determined by the resonant frequency of the very high Q series LC network in the feedback loop. This network is made up of the 4–30 pF variable capacitor and quartz crystal which operates in the parallel mode. The variable capacitor provides a fine frequency adjustment control. Feedback is sufficiently large to assure normal circuit operation almost completely independent of transistor gain. The large amount of feedback drives the collector from cutoff to saturation, making a square

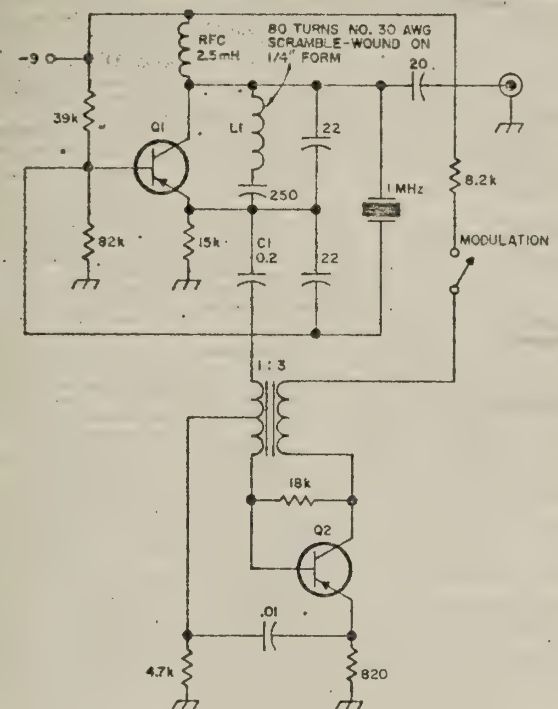


Fig. 65. This crystal controlled oscillator provides very distinctive markers up to 30 MHz. The modulation frequency is approximately 1000 Hz, but by changing the value of C1 it may be changed slightly. Q1 is a 2N384, 2N1742, 2N2362, 2N2084, TIM10, GE-9 or HEP-2; Q2 is a 2N2613, 2N2953, 2N1303, SK3004, GE-2 or HEP-254.

wave available at the output. A sine wave is developed across a tunable high Q tank and is also available at the output.

Modulated band edge marker

The crystal controlled calibration oscillator shown in Fig. 65 is especially useful for band edge marking and providing distinctive crystal controlled markers up to 30 MHz. To assist in identification of the marker, particularly at the higher frequencies where the harmonics are quite weak, the note may be modulated by simply turning on the audio oscillator. The modulation frequency of this unit is about 1000 Hz, but it may be changed slightly by changing the value of C1. If the oscillator fails to oscillate when power is applied, reverse the connections on one side of the transformer. At the lower frequencies the output of the calibrator may be coupled into the antenna circuit of the receiver by inductive coupling, but on 15 and 10 meters, a direct connection to the antenna terminals may be necessary to obtain sufficient output. Although this circuit is designed for a one MHz crystal, other crystal frequencies may be accommodated by changing the number of turns in L1.

The two meter band edge marker illustrated in Fig. 66 provides very strong harmonics of an 18 or 24 MHz crystal on 144 MHz; when a sensitive converter is used on 432, harmonics may also be heard on this band. This circuit will oscillate with crystals throughout the 18 to 24 MHz region, so it may be used as a marker at almost any VHF frequency. The use of a 20 MHz crystal for example would be very useful for marking the lower edge of the amateur 220 and 420 MHz bands. If a modulated marker is desired, the audio oscillator (Q2) of Fig. 65 may be coupled into the base of the oscillator through a 0.2 μF capacitor.

Sweep frequency generator FIG 67

More and more hams are finding out that the sweep generator is one of the most useful test instruments to have around the shack. It may be used to align communications receivers, VHF converters, to plot response curves and to check bandwidth. The circuit illustrated in Fig. 67 is a very simple unit that may be used at any spot frequency between 100 kHz and 60 MHz. Although a three range bandswitch is shown in the drawing, it may be omitted if only one spot frequency is required (455 kHz for example).

The sweep generator shown here consists of a single unijunction transistor sawtooth generator which provides the sweeping signal for the oscilloscope and the fixed tuned rf oscillator. The output of the sawtooth generator is connected across a 56 pF varicap in the oscillator tank which varies the frequency of the oscillator in step with the scope trace. The sweeping frequency may

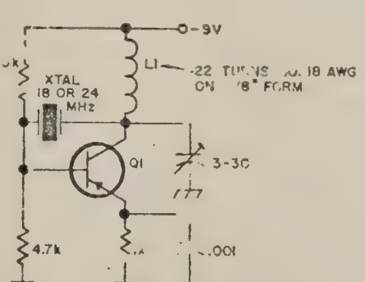


Fig. 66. This two meter band edge marker provides useful harmonics up to several hundred MHz with 18 or 24 MHz crystals. If modulation is desired, the audio oscillator from the circuit of Fig. 65 may be coupled into the base. Q1 is a 2N1745, 2N2362, or HEP-2.

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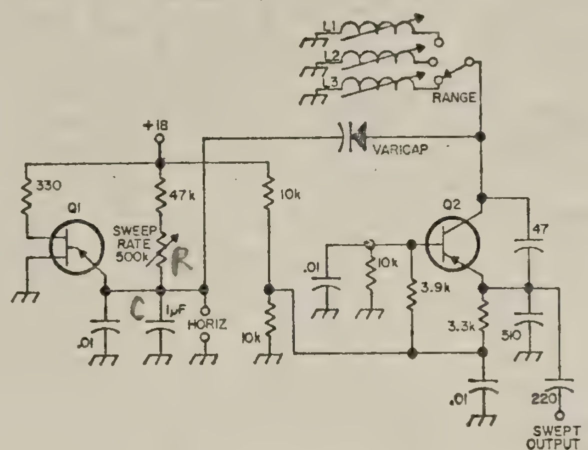


Fig. 67. A sweep frequency generator is a very handy gadget, but many times the commercial units are more complicated than the ham requires. This simple sweeper may be used at any spot frequency between 100 kHz and 60 MHz by simply using the coils listed in Table 2. By using a three position range switch, the three most popular frequencies may be used, such as 455 kHz, 1600 kHz and 10.7 MHz. Q1 is a 2N1671, 2N2160, 2N2646, 2N3480, or HEP-310; Q2 is a 2N741, 2N1747, 2N2188, GE-9 or HEP-2. The varicap is a 56 pF capacitance diode such as the 1N955 or TRW V56.

Motorola EPICAP

be varied between 5 and 30 sweeps per second with the 500k sweep rate control.

The center frequency of the swept output is determined by the coils that are connected across the range switch. A listing of coils for all frequencies from 65 kHz to 60 MHz is listed in Table 2. When using this generator, always use the slowest sweep speed that will provide a usable trace on your oscilloscope. The slower the sweep speed, the more accurate the reproduction of the response curve; if the generator is swept at too fast a rate, the resonant circuit being swept may ring and distort the curve.

Table 2
COILS

Frequency		Miller No.
65 kHz—140 kHz		9007
95 kHz—190 kHz		9006
150 kHz—300 kHz		9005
190 kHz—550 kHz	—455 KC I.F.	9004
380 kHz—1000 kHz		9003
700 kHz—1.8 MHz		9002
1.4 MHz—3.7 MHz		9001
3.7 MHz—4.7 MHz		4508
4.7 MHz—5.9 MHz		4507
5.9 MHz—7.5 MHz		4506
7.5 MHz—10 MHz		4505
10 MHz—14 MHz	—10.7 KC I F	4504
14 MHz—18 MHz		4503
18 MHz—23 MHz		4502
23 MHz—29 MHz		4304
29 MHz—36 MHz		4303
36 MHz—45 MHz		4302
45 MHz—60 MHz		4301

Sawtooth generator

Sawtooth generators are very useful in many measurements and their circuitry may be greatly simplified by the use of field effect transistors. When conventional junction transistors are used for this purpose, complex feedback networks and methods of compensation must be used to generate a linear voltage ramp. The output waveform of the sawtooth generator shown in Fig. 68 is linear within 2% and may be adjusted from 1 kHz to 3 kHz by the center frequency control. The thermistor R1 provides temperature stability and circuit loading is reduced by the use of a source follower at the output.

Square wave generator

The square waves available from most inexpensive signal generators deteriorate pretty badly at the higher frequencies. In

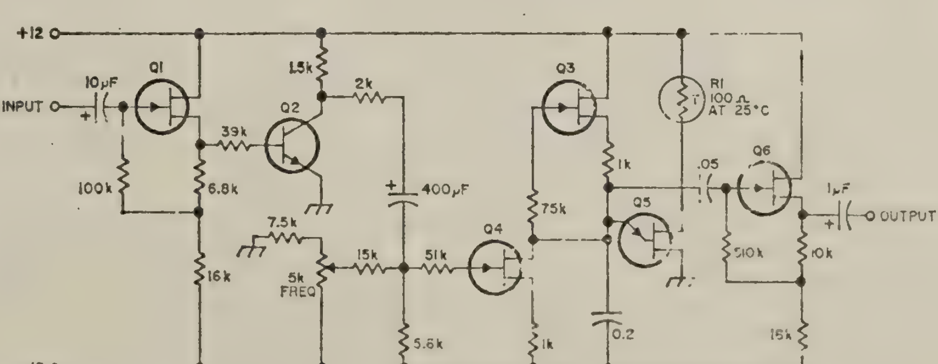
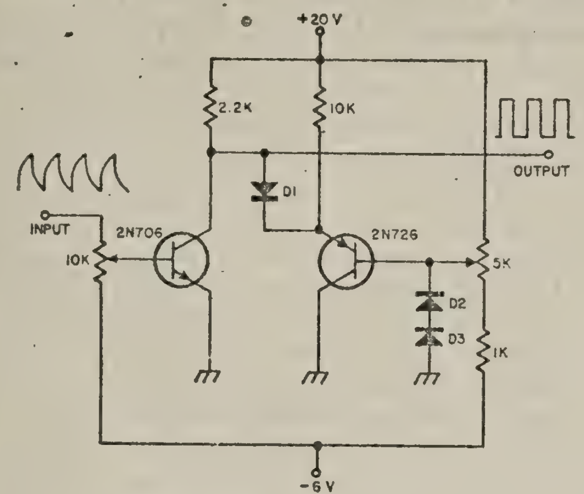


Fig. 68. This simple sawtooth generator is linear within 2% and may be adjusted from 1 kHz to 3 kHz with the center frequency control. Q1, Q3, Q4 and Q6 are FET's such as the 2N3819, 2N3820, TIS34, MPF.05 or HEP 801; Q2 is a 2N388, 2N2926, 2N3391, SK3011 or HEP-54. Q5 is a 2N1671, 2N2160, 2N2646, 2N3480 or HEP-310.



ALL DIODES GENERAL PURPOSE SILICON

Fig. 69. The square wave output of many inexpensive signal generators deteriorates quite badly at high frequencies, but this circuit will square them off again. The diodes may be any inexpensive computer type such as the 1N914.

addition, the output amplitude often varies with frequency. The addition of the simple circuit illustrated in **Fig. 69** to the output of the signal generator will provide a good square wave output with an amplitude that does not vary with frequency.

In this circuit, Q1 is an inverter, while Q2 and the 5 kilohm amplitude control pot provide a variable clamp voltage for the output. In most cases, the necessary positive and negative voltages may be obtained from the signal generator's internal supply. The input potentiometer should be adjusted for best output waveform. Once this control is set, it will not change unless the signal generator varies significantly. Ideally, this control should be adjusted so that the input signal provides a voltage swing at the base of Q1 from +3 to −5 volts. The silicon diode D1 provides protection for the emitter-base junction of Q2. Diodes D2 and D3 prevent the clamp voltage from reverse biasing Q1.

Capacitance meter

The direct reading capacitance meter illustrated in **Fig. 70** has four direct reading capacitance ranges from zero to 0.1 μF. Although electrolytic capacitors cannot be measured with this circuit, any type of non-electrolytic capacitor may be checked. In fact, it works as well with variable capacitors as with fixed; the meter deflection will follow the capacitance change as it is tuned. The four direct reading ranges are 0-200 pF, 0-1000 pF, 0-0.01 μF and 0-0.1 μF. The lowest capacitance which may be accurately read is 4 pF, but 2 pF may be estimated quite easily.

In this circuit transistors Q1 and Q2 are connected in a conventional free running multivibrator. The output from the collector of Q2 is a constant amplitude square wave whose frequency is determined by the values of the resistors and capacitors connected across the base circuit of Q1 and Q2. The square wave output from the collector of Q2 is coupled through the unknown capacitor connected across the test terminals to the metering circuit consisting of the 1N34A diode, the potentiometers and the dc microammeter. For any given square wave frequency, the deflection of the meter will be directly proportional to the capacitance across the test terminals. For example, if a precision 100 pF capacitor is connected across the test terminals and the calibration pot is set for full scale deflection, the

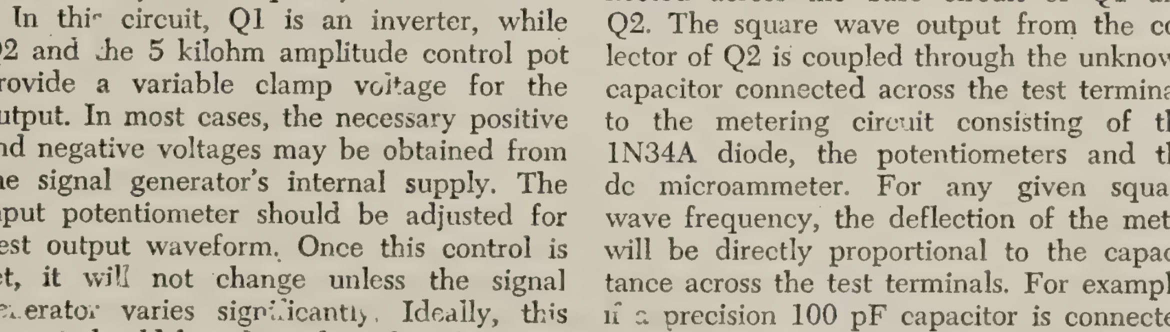


Fig. 70. Although this capacitance meter will not measure electrolytic capacitors, it will measure any other type from zero to 0.1 μF with reasonable accuracy. On the lower end 4 pF can be read accurately and 2 pF easily estimated. Transistors Q1 and Q2 are 2N168, 2N1605 2N2926, SK3011 or HEP-54. The meter is a 0-50 microampere unit and the range switch a Centralab PA1021.

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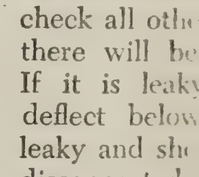

Fig. 71 Thi… 1200 megohm… Upp… 5K pot. Q1 is … Q2 is … 2N706…

meter will have a range from zero to 100 pF. The response of this circuit is essentially linear, so 50 pF would provide half-scale deflection, 1/10 scale deflection would indicate 10 pf, etcetera. To change the range, the frequency of the multivibrator is changed by choosing new values of resistance and capacitance in the base circuits of Q1 and Q2.

Calibration of the direct reading capacitance meter requires the use of four accurately known capacitors—0.1 µF, 0.01 µF, 1000 pF and 200 pF. These capacitors should be very carefully chosen because the accuracy of the completed meter depends on the tolerance of the calibration capacitors. In addition, it is essential that they are not leaky; if they are, the calibration will not be accurate. To calibrate the meter, set the range switch in the appropriate position, connect the respective calibrating capacitor across the test terminals, and adjust the calibrating potentiometer for full scale deflection of the meter.

Although electrolytic capacitors cannot be measured on this instrument, it will check all other types. If a capacitor is open, there will be no deflection on any range. If it is leaky or shorted, the meter will deflect below zero. It is imperative that leaky and shorted capacitors be immediately disconnected from the test terminals to prevent damage to the diode and microammeter.

High impedance scope probe

Most oscilloscope probes are unsuitable for use with high impedance circuits because they severely load them down. By using the high input impedance characteristics of the field effect transistor, an extremely high impedance probe may be produced. The circuit shown in Fig. 71 uses

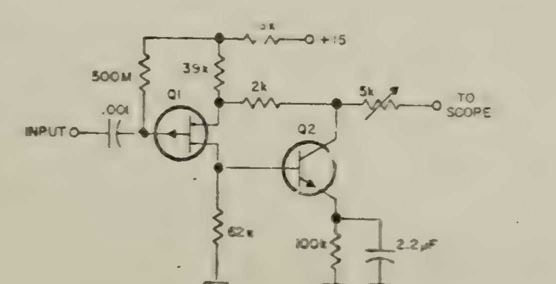

Fig. 71. This high impedance probe provides about 1200 megohms input impedance with unity gain. Upper frequency equalization is provided by the 5K pot. Q1 is a U112, 2N2607, 2N3360 or TIM12; Q2 is a 2N706, 2N708, 2N2926, 2N3391 or HEP-50.

the bootstrap action of the 500 megohm resistor to raise the input impedance of the circuit to 1200 megohms. The 2 kilohm feedback resistor from the collector of Q2 maintains unity gain while the 5k potentiometer provides circuit equalization. The rise time of this circuit is extremely fast, typically less than half a microsecond. In addition, it can handle up to two volt signals (peak to peak) without distortion.

Microammeter amplifier

The sensitive microammeter amplifier illustrated in Fig. 72 may be adjusted from approximately 2 µA to 100 µA full scale deflection each side of zero. The input impedance of this circuit is 60 kilohms at 2 µA sensitivity and 2.5 kilohms at 100 µA sensitivity; total battery drain is 1.5 mA.

In this circuit a differential amplifier with degenerative biasing and collector meter feed provide a satisfactory compromise between sensitivity and stability. The transistors used in this circuit should be chosen for high current gain throughout the emitter current range; in addition, they should exhibit very low leakage currents. The 50 ohm null potentiometer is used to compensate for any differences in the base-emitter voltage of the two transistors. The balance pot is used to compensate for dif-

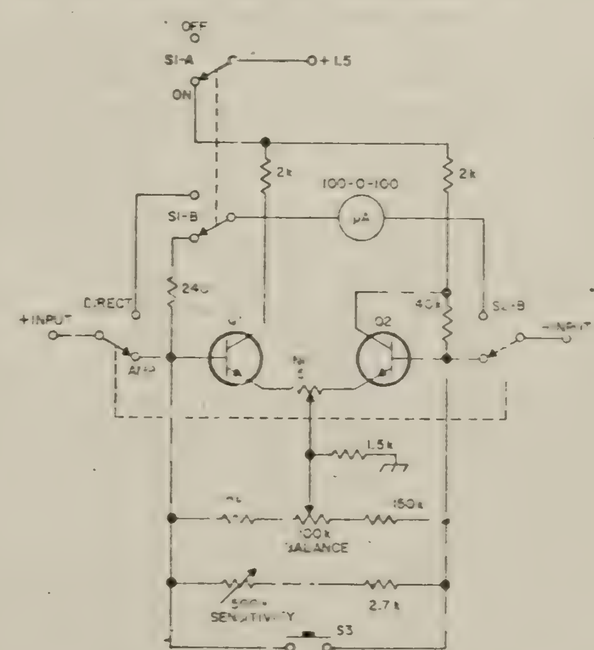

Fig. 72. This very sensitive microammeter amplifier may be adjusted from 2 µA to 100 µA each side of zero; the input impedance varies from 60K at 2 µA to 2.5K at 100 µA. Transistors Q1 and Q2 are 2N930, GE-10 or HEP-50.

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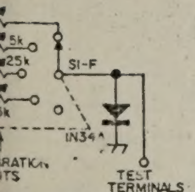

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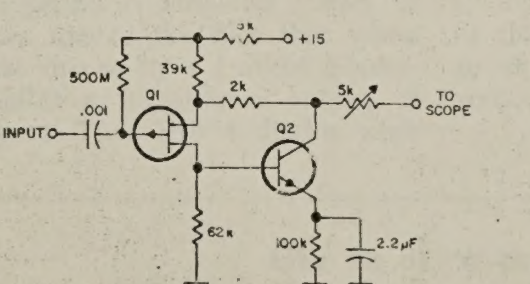

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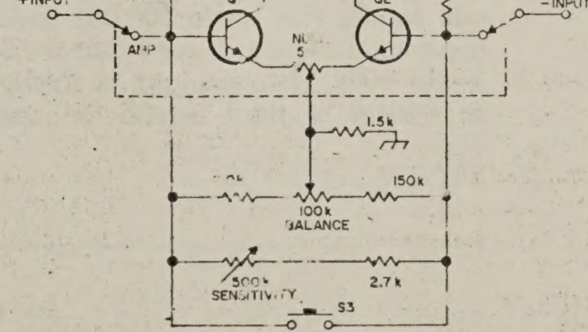

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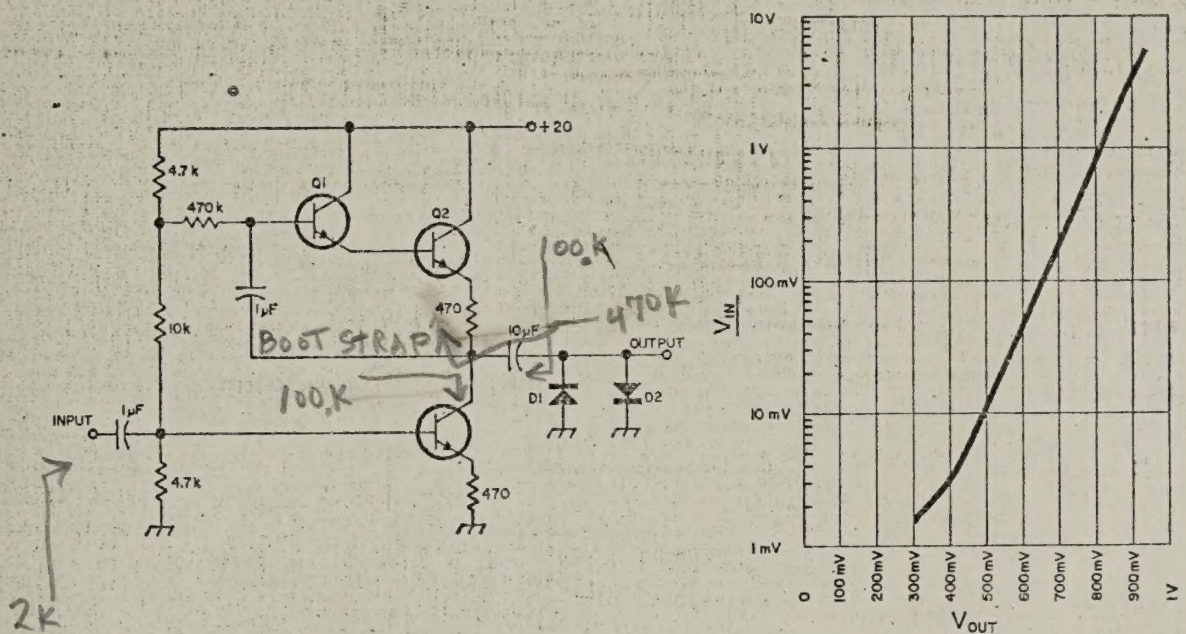

Fig. 73. This logarithmic amplifier makes use of the fact that when two back to back diodes are driven by a current generator, they exhibit a logarithmic output of the input signal. With the circuit constants shown, this amplifier follows a nearly perfect logarithmic curve over a 60 dB range; selected diodes will increase this to 80 dB. Q1, Q2 and Q3 are 2N2924, SK3019, GE-10 or HEP-54; D1 and D2 are 1N914.

ferences in components and transistor current gain.

The null adjustment is made with S3 depressed, S1 on and S2 on "amplifier." The balance is adjusted with S1 on and S2 on "direct." Although the null is completely independent of the balance adjustment, the balance must be changed each time the null is changed. For improved sensitivity a 50-0-50 μA meter may be used instead of the 100 μA movement, but at the cost of reduced stability.

Logarithmic amplifier

Logarithmic amplifiers are very useful for making measurements that require very large changes in input voltage or current. The 60 dB logarithmic amplifier shown in Fig. 73 makes use of the fact that when two diodes are driven by a current generator, they exhibit a logarithmic output of the input signal. Two 1N914 diodes were chosen for this amplifier because they follow an almost perfect logarithmic curve over a 60 dB range; by selecting diodes, it is possible to obtain the same type of curve over an 80 dB spread.

To insure that load changes are not affected by Q1, its output impedance must be extremely high; to cover the 60 dB range for example, the impedance must exceed 100 kilohms. This is accomplished by the Darlington pair, Q1 and Q2 and the 470 ohm resistor in the emitter of Q2. If Q3 is maintained in its linear range, its ac collector current is completely independent of the collector load. Hence, it presents a very high output impedance to the diodes; greater than 100,000 ohms with almost any silicon planar transistor. The input impedance of this circuit is approximately 2000 ohms, so it may be driven from a variety of sources.

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